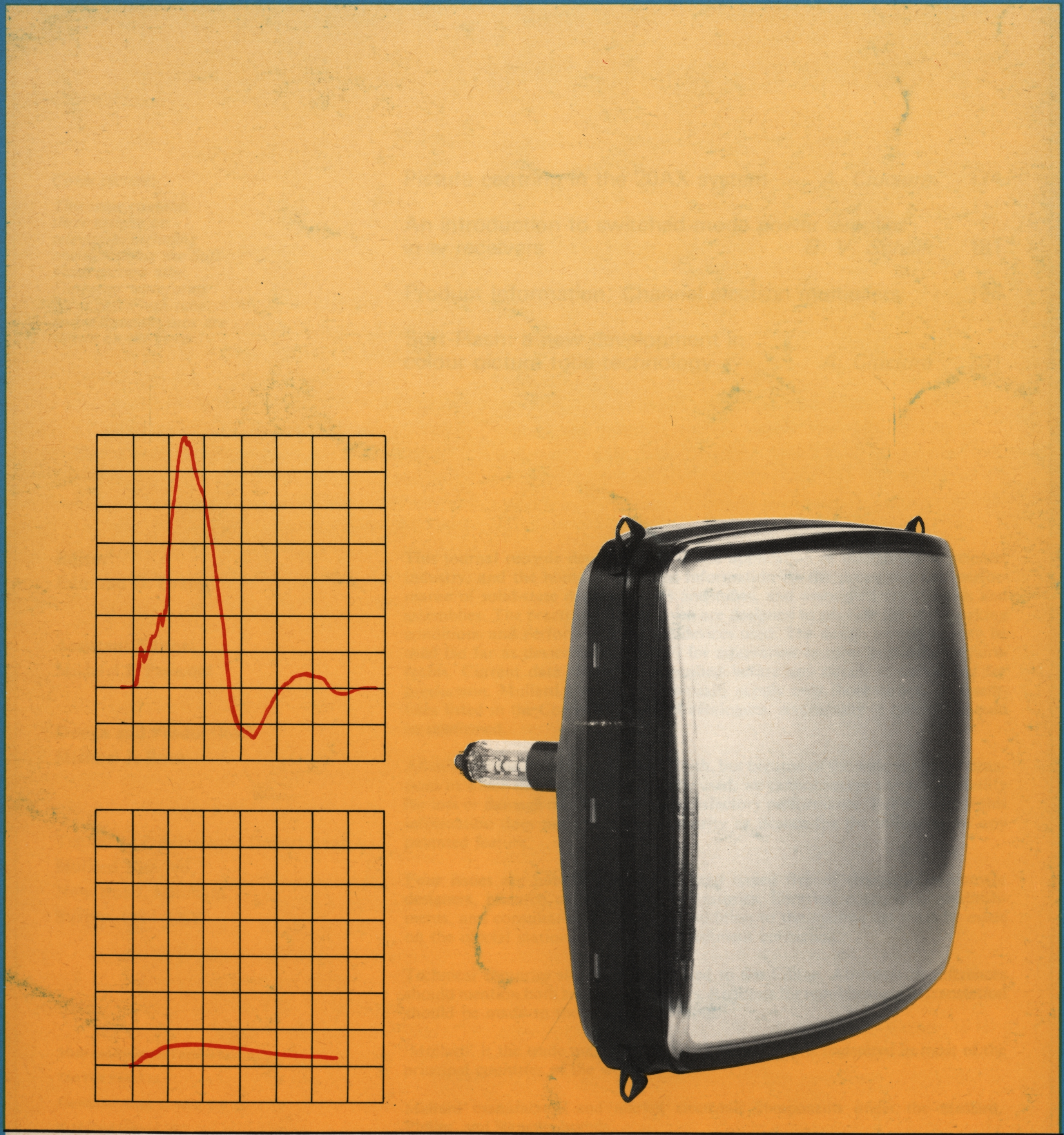
Mullard

Technical Communications

Volume 14 Number 135
July 1977



- Picture centring in the 20AX system
- An introduction to SMPS in tv receivers
- Channel electron multipliers
- Soft-Flash: a new development in colour picture tube technology

Mullard Technical Communications

ISSN 0027—3139

Volume 14 Number 135 July 1977

Cover picture:

This issue contains three articles on television, including one describing the Soft-Flash picture tube. Flashover waveforms for a Soft-Flash tube and a standard tube are shown on the cover.

Picture centring in the 20AX system *A. Ciuciura* 174

An introduction to switched-mode power supplies in tv receivers *G. V. Schaik* 181

Product Information: Channel electron multipliers 196

Soft-Flash: a new development in colour picture tube technology *A. Ciuciura* 201

Editor:
Edmund G. Evans

Assistant Editor:
Michael J. Prescott

Design and Production:
Michael J. Rose

Technical Publications Department
Mullard Limited,
New Road, Mitcham,
Surrey, CR4 4XY

Made and Printed in England by
The Yale Press Ltd.,
Carmichael Road, London, S.E.25.

© Mullard Limited 1977

This journal records developments in the application of electronics in science, industry, and the home. It contains information on the properties and performance of solid-state devices, valves and tubes, and associated components and assemblies. The practical circuits given are designed to provide the best working conditions and performance for the devices used. The design principles may be used (so far as device ratings allow) for adaptation to meet particular requirements. Current data should be consulted before any circuit is considered for production. Mullard Limited can normally supply only those devices or assemblies listed in the Company's current catalogues. Availability in quantity should be confirmed.

All information is published in good faith, but because of the wide variety of situations in which this information may be used, we cannot accept any responsibility for loss or damage resulting from unsatisfactory performance. Publication of this information does not imply any authority or licence for the utilisation of any patented feature.

Four issues are published each year and issued free of charge to equipment designers, research organisations, laboratories, libraries, educational establishments, and consultants. Requests for addition to the mailing list must be made on the official stationery of the establishment concerned.

Technical enquiries should be addressed to the Editor. Abstracts or references should mention both source and authors. Requests for republication or translation should be made to the Editor in writing.

'Mullard' is the trade mark of Mullard Limited and is registered in most of the principal countries of the world.

Mullard manufacture and market electronic components under the **Mullard, Philips,** and **Signetics** brands.

Mullard Limited, Mullard House, Torrington Place, London WC1E 7HD Tel: 01-580 6633 Telex: 264341

Picture centring in the 20AX system

A. Ciuciura
Mullard Application Laboratory

This article describes the centring of picture information on the screen in the 20AX television system. The reduced tolerance on the positioning of the undeflected beam in this system has made it possible to omit the specialised picture centring and use other controls to adjust the centring. The interrelated factors affecting the picture centring of 20AX are discussed, and their correction described; a set of recommendations for adjustment procedure and a range of diagnostic tests are given.

INTRODUCTION

One of the features of the 20AX system is a very tight tolerance on the positioning of the undeflected beam at the centre of the screen; this tolerance, about 5mm, compares well with 15mm for 90° tubes and 12mm for earlier 110° tubes. In most applications it is now possible to omit the picture centring circuit of the receiver, a long-standing wish of many circuit designers. If this is done, the only additional requirement is an extension in the range of the phase-shift control of the line oscillator (by about ±0.6μs) to provide an accurate positioning of picture information on the screen. With the length of flyback time used at present, the loss of picture information resulting from use of the phase control as a picture shift control is a maximum of 2% of the raster width, and this loss is considered acceptable in the majority of applications. If it is required to display 100% of picture information with accurate picture centring, then a picture centring circuit must be used in conjunction with the phase control, although the range of the centring circuit can be reduced for 20AX in comparison with other systems.

In a well-adjusted receiver, the picture centring circuit should be used primarily to compensate for picture-tube tolerances. However, in practice it is used to correct a number of other effects, and in this respect its absence can create new problems which require some care in the design and production of the receiver. In return, the omission of the centring circuit can give better performance, and offers the advantages of lower cost, improved reliability, and saving of space.

This article discusses the factors involved in picture centring and their complex interdependence. A set of recommendations is evolved, to assist with problems arising from the absence of the picture centring circuit in the 20AX system.

TERMS USED IN PICTURE CENTRING

Picture centring is commonly thought of as the process of aligning the picture information (video) with the screen area of the picture tube, and although this is correct in principle, in practice the situation is more complex. Even in the simple model shown in Fig. 1 it can be seen that the picture centring involves the alignment of four points; these points are defined below for convenience.

Geometric centre of the useful screen area

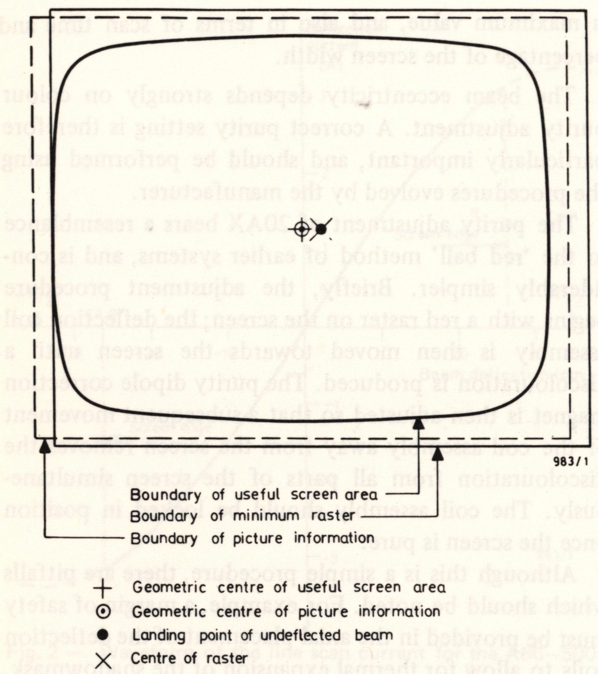

983/1
Boundary of useful screen area
Boundary of minimum raster
Boundary of picture information
+ Geometric centre of useful screen area
⊙ Geometric centre of picture information
● Landing point of undeflected beam
× Centre of raster

Fig. 1 — Elements of picture centring

The useful screen area is the area of the television picture tube which is illuminated during normal operation; its minimum dimension is invariably quoted in the published data. The centre of this area is a vital point in the investigation of picture centring problems. For aesthetic reasons the Mullard range of tubes does not carry any identification marks to assist with the location of this point, but it may be located at the crossing of the two diagonals; care must be taken to avoid parallax errors.

Landing point of undeflected beam

The landing point is that point in the useful screen area where the undeflected electron beam strikes the screen (after colour purity, static convergence, and horizontal centre-line adjustments have been made). The distance between this point and the geometric centre of the useful screen area is called the beam eccentricity or centring error of the picture tube, and is usually quoted in the published data.

The beam landing point must be located with great care in order to avoid errors arising from any disturbance or incorrect setting of the AT1081 multipole unit, and to avoid any possibility of a spot burn. Location of the point can be done in a television receiver provided that the picture tube can be cut-off (while maintaining the correct V_{g_2} voltages), and that the space around the tube is free from strong magnetic fields. Leads from the deflection coils should be connected to another deflection coil assembly, well away from the tube, to give the same circuit loading while removing any beam deflection. If a very precise determination is required, the effects of the Earth's magnetic field must be taken into consideration by rotating the receiver 180° about the tube axis after one measurement, degaussing, and repeating the measurement. The true beam landing point then lies halfway between the points determined before and after rotation.

Centre of the raster area

This point is the geometric centre of the area scanned by the electron beams; this area is usually taken to be a complete raster, with blanking removed, which is entirely visible on the screen. The centre of the raster area, unless deliberate adjustments are made, is not coincident with the beam landing point discussed above. The main cause of this discrepancy is raster non-linearity, although there is a slight contribution due to the shape of the flyback waveform. This second quantity may be a function of the line timebase circuit (transistor and thyristor circuits producing different effects) and the loading of the line output transformer (by the high-voltage supply for example).

Geometric centre of picture information

This point is defined as the geometric centre of the video information being received. During adjustment of a receiver it is customary to display a test card, and most of these have a well-defined centre of picture information; location of the displayed information in relation to the raster depends on relative timing. In modern receivers, the shifting of the picture information with respect to the raster is controlled in the horizontal direction by the phase shift of the line oscillator. No equivalent control exists in the vertical direction.

PICTURE CENTRING ADJUSTMENT MECHANISMS

Three separate controls are used to adjust picture centring; these are described below.

Picture shift control

The picture shift circuit is usually used to correct the beam eccentricity of the picture tube; as the eccentricity of 20AX tubes is very small, this correction may be performed satisfactorily by the phase shift circuit (see later) in most applications.

Linearity control

The linearity control circuit is used to improve raster linearity, and this brings the centre of the raster closer to the landing point of the undeflected beam.

Phase shift control

The phase shift circuit is primarily intended for the alignment of picture information with the raster, in the horizontal direction.

It will be noted that each of these control circuits performs the same function, that is the changing of the position of the picture on the screen. When the raster overscans the useful screen area, as is usually the case, it is not immediately obvious which control is in operation, and injudicious use of the control circuits can easily lead to confusion and consequent maladjustment of the receiver. The following section considers in detail the parameters affecting picture centring in the 20AX system, and how centring faults may be corrected. A description of an orderly adjustment procedure is given later.

PARAMETERS AFFECTING PICTURE CENTRING

The factors affecting picture centring in the 20AX system may be separated into three categories, which are listed below:

1) beam eccentricity, and the effects of purity and other adjustments on it;
2) raster amplitude, and the effects of asymmetric blanking and non-linearity of the line-scan current;
3) phase discrepancy between the video information and the line timebase.

Each of these factors has its own set of tolerances and definite limits within which the results are considered to be acceptable. Although the three categories are described separately below, they are interrelated to such an extent that a cursory examination of a faulty receiver can easily lead to a mistaken diagnosis of the problem. The description given below is in terms of line scan and horizontal displacements, but most of the arguments apply equally well to the vertical direction. However, vertical adjustments are less complex and, in general, do not present much difficulty.

Beam eccentricity

Beam eccentricity, or picture tube centrality, is defined in relation to the geometric centre of the useful screen area. Normally the published data quotes this tolerance as a maximum centring error in any direction after colour purity, static convergence, and horizontal centre-line corrections have been correctly adjusted. Table 1 gives these tolerances for four sizes of 20AX tubes as a maximum value, and also in terms of scan time and percentage of the screen width.

The beam eccentricity depends strongly on colour purity adjustment. A correct purity setting is therefore particularly important, and should be performed using the procedures evolved by the manufacturer.

The purity adjustment of 20AX bears a resemblance to the 'red ball' method of earlier systems, and is considerably simpler. Briefly, the adjustment procedure begins with a red raster on the screen; the deflection coil assembly is then moved towards the screen until a discolouration is produced. The purity dipole correction magnet is then adjusted so that a subsequent movement of the coil assembly away from the screen removes the discolouration from all parts of the screen simultaneously. The coil assembly should be locked in position once the screen is pure.

Although this is a simple procedure, there are pitfalls which should be noted. For example, a margin of safety must be provided in the axial placement of the deflection coils to allow for thermal expansion of the shadowmask, and although there are many settings of the deflection coils and the purity magnet giving a pure raster, only one setting gives correct picture centrality and a sufficient margin for shadowmask expansion.

Raster

Waveform of the line scan current

In order to produce a linear display on the screen, the line scan current must be 'S' corrected, the resulting current waveform closely resembling a linear portion of a sinewave curve. In the 20AX system the amount of correction is larger than for earlier tubes, and a typical scan current waveform is shown in Fig. 2. The continuous curve represents the edge-to-edge deflection current of an A66-500X picture tube, and the dashed continuation gives the normal sinewave curve.

TABLE 1

Minimum Screen Width and Beam Eccentricity of 20AX Picture Tubes

Tube type	Minimum screen width (mm)	Beam eccentricity		
		Maximum (mm)	Percentage of screen width	Period (μs)*
A66-500X	518	5.0	0.97	0.506
A56-500X	444	4.5	1.01	0.531
A51-500X	404	4.5	1.11	0.584
A47-500X	369	4.0	1.08	0.568

*Assuming a flyback time of 11.6μs

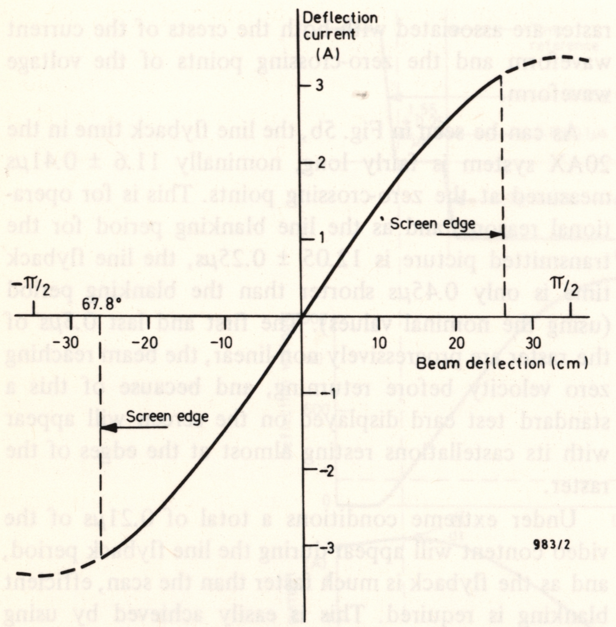


Fig. 2 — Waveform of the line scan current for the A66–500X picture tube

As the waveform of the deflection current shows, the scan sensitivity is a function of the deflection angle; the sensitivity of the system increases with deflection. Although the variation is continuous (inversely proportional to the cosine of the deflection angle), for many purposes it can be assumed that a change in the centre of the screen produces a doubled effect at the edges. As an illustration, consider a picture tube with zero beam eccentricity and a linear raster which scans the tube accurately from edge to edge. If by some means the beam eccentricity is changed to +5mm, say (the beam landing a maximum of 5mm to the right of the geometric centre), then the whole raster is shifted to the right. The right edge of the screen is overscanned while the left edge shows a blank margin, and the width of the margin is not a maximum of 5mm, as would be the case in a system with a linear scan current, but up to 10mm. If overscanning is corrected by reducing the raster amplitude, so that the right-hand end corresponds with the right edge of the screen, the blank margin on the left is increased to up to 20mm. In this way the non-linearity of the scan current can lead to the conclusion that the beam eccentricity is outside the ±5mm limit. However, if the raster is examined closely the left side will be found to be cramped, and when this is corrected by reducing the amount of linearity correction, and the right-hand end of the raster is aligned with the edge of the screen, the blank margin on the left is reduced to about 10mm.

The effects described above indicate that the amount of linearity correction is a function of beam eccentricity, and also that an incorrectly adjusted raster can give the impression of excessive beam eccentricity.

Linearity correction

The line linearity control compensates for the resistive voltage component in the deflection circuit, and as a result its effect is most pronounced on the left-hand side of the raster. If the beam eccentricity is such that the undeflected beam lands to the right of the geometric centre, the left side of the raster is moved towards the centre of the screen, into the area of lower sensitivity, and this provides some automatic raster correction. Less correction is therefore required from the line linearity control coil. If the undeflected beam lands to the left of the geometric centre (a beam eccentricity of −5mm say), however, a large amount of linearity correction is required.

An experimental graph showing the relationship between line linearity correction voltage and beam eccentricity is shown in Fig. 3. This graph is a straight line, with a gradient of about 1.6V per mm of beam eccentricity. It should be noted that the mean value of the line linearity correction voltage depends on the line convergence circuit used, and to a lesser extent on the layout of the circuit. For example, removal of the line balance parabola part of the convergence circuit produces a drop of about 5V in the required mean value of the correction voltage.

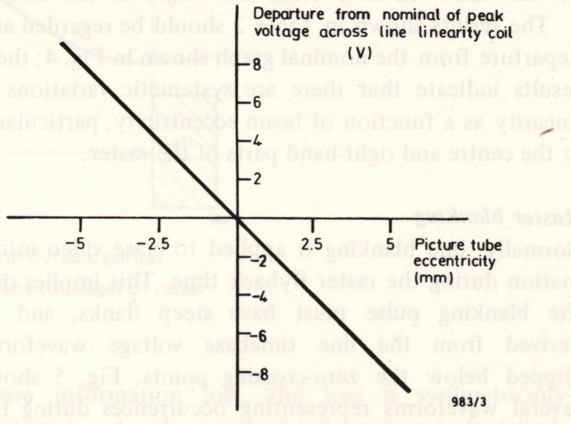


Fig. 3 — Line linearity correction voltage plotted against beam eccentricity

One result of the variation in the line linearity correction voltage is variation of scan amplitude, which is inversely proportional to the correction voltage. A raster which is correctly adjusted on a tube with zero beam eccentricity will be overscanned by about 16mm on a tube with a beam eccentricity of −5mm (assuming correct linearity adjustment in each case).

The differences in line linearity correction should also affect the 'S' correction, although this is automatically compensated to a large extent by adjustment of the line linearity coil. However, there are residual effects which are reflected in changes of raster linearity with variations

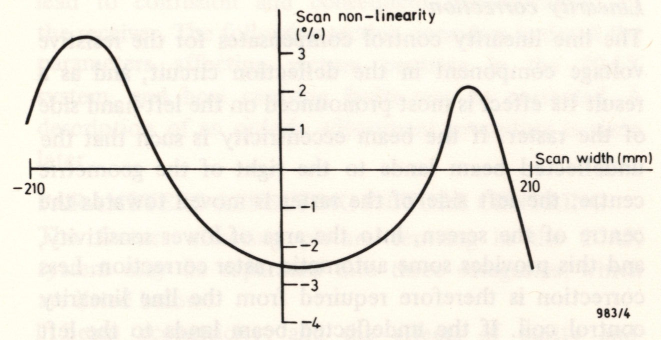


Fig. 4 — Line scan non-linearity for an A56–500X picture tube with zero beam eccentricity

in beam eccentricity. Fig. 4 gives a graph of line scan non-linearity for an A56-500X tube having zero beam eccentricity, and Table 2 gives values of scan non-linearity at the left edge, centre, and right edge of the raster for various values of beam eccentricity.

The raster non-linearity was measured by setting the centre of a cross-hatch pattern to the geometric centre of the screen, and adjusting the edges of the pattern to be the same distance from the edges of the screen.

The results shown in Table 2 should be regarded as a departure from the nominal graph shown in Fig. 4; these results indicate that there are systematic variations in linearity as a function of beam eccentricity, particularly at the centre and right-hand parts of the raster.

Raster blanking

Normally, line blanking is applied to erase video information during the raster flyback time. This implies that the blanking pulse must have steep flanks, and be derived from the line timebase voltage waveform, clipped below the zero-crossing points. Fig. 5 shows several waveforms representing occurrences during the flyback period. Fig. 5a gives the video waveform defined by the transmission system; Figs. 5b and 5c show simplified current and voltage waveforms; and Fig. 5d shows the corresponding appearance of the edges of the raster. It should be noted that the furthest excursions of the raster are associated with both the crests of the current waveform and the zero-crossing points of the voltage waveform.

As can be seen in Fig. 5b, the line flyback time in the 20AX system is fairly long, nominally 11.6 ± 0.41μs measured at the zero-crossing points. This is for operational reasons, and as the line blanking period for the transmitted picture is 12.05 ± 0.25μs, the line flyback time is only 0.45μs shorter than the blanking period (using the nominal values). The first and last 0.5μs of the raster are progressively non-linear, the beam reaching zero velocity before returning, and because of this a standard test card displayed on the screen will appear with its castellations resting almost at the edges of the raster.

Under extreme conditions a total of 0.21μs of the video content will appear during the line flyback period, and as the flyback is much faster than the scan, efficient blanking is required. This is easily achieved by using readily available ICs, but there are possible problems. Any delay or lengthening of the blanking pulse can produce asymmetric blanking, obliterating part of the raster. In this way a properly centred raster can give the appearance of any asymmetrically placed raster, leading to the erroneous conclusion that the beam eccentricity is greater than it should be.

Phase Discrepancy

In the preceding sections it has been assumed that the phase relationship between the video information and the line timebase is ideal. This is not always the case, and even a correctly adjusted circuit can have some variation in phase with time, but with efficient blanking and proper overscan such variation is not noticeable by the viewer. However, another dimension is added by the omission of the picture centring circuit, as the phase control must be used to centre the picture on the screen. Application of phase control when the picture information practically fills the raster as described earlier may cause some blanking of picture information on one side of the screen, and leave a small part of the raster blank on the other side. This is the situation illustrated in Fig. 1, showing the landing point of the undeflected beam to be to the right of the geometric centre of the useful screen area. If the raster is adjusted for good linearity, then it will scan the tube for an equal distance to either side of the landing point, and with the scan amplitude adjusted so that the left-hand end of the raster coincides with the left edge of the screen, the right edge is overscanned by about twice the beam eccentricity.

A normal picture could be displayed on this screen with zero phase shift and the viewer would not be aware of the small amount of eccentricity (about 1%). A test card, however, is designed to have symmetrical features and it would immediately be noticed that a particular vertical line coincident with the right edge of the screen was 10mm away from the left edge in the limt case. It is possible to correct for this error by phase-shifting the picture information by the same amount as the beam eccentricity and in the opposite direction; although this method does not alter the amount of picture information lost, the loss is symmetrically distributed and only marginal information is forfeited.

As mentioned above, the amount of phase shift required for picture centring is equal to the beam eccentricity, which is shown in Table 1. It can be seen that it does not exceed ±0.58μs for a 20AX tube, representing a maximum of 1.1% of picture width. If all the receivers being produced on an assembly line are to show the same display, the overscan on each receiver must be 2% of the complete scan.

The requirements for phase shift and overscan described above are additional to the normal overscan requirements to cover spreads in component values, and these new requirements may be added to the normal ones in r.m.s. fashion. If this is done, it is found that a standard overscan of 6% should be increased to about 6.5%, which is a very small change in practical terms. However, if the TBA920 line oscillator combination IC (see published data) is used, its phase tolerance of ±0.7μs must be included, and in this case the overall adjustment of the phase control should be at least 1.1μs.

TABLE 2

Line Scan Non-linearity as a Function of Beam Eccentricity

Beam eccentricity (mm)	Raster non-linearity (%) Left edge	Centre	Right edge
−5.0	0.00	−3.0	1.7
−2.5	0.50	−2.8	0.5
0.0	1.24	−2.5	−2.7
+2.5	−1.21	−2.0	−3.5
+5.0	−1.60	−1.6	−5.9

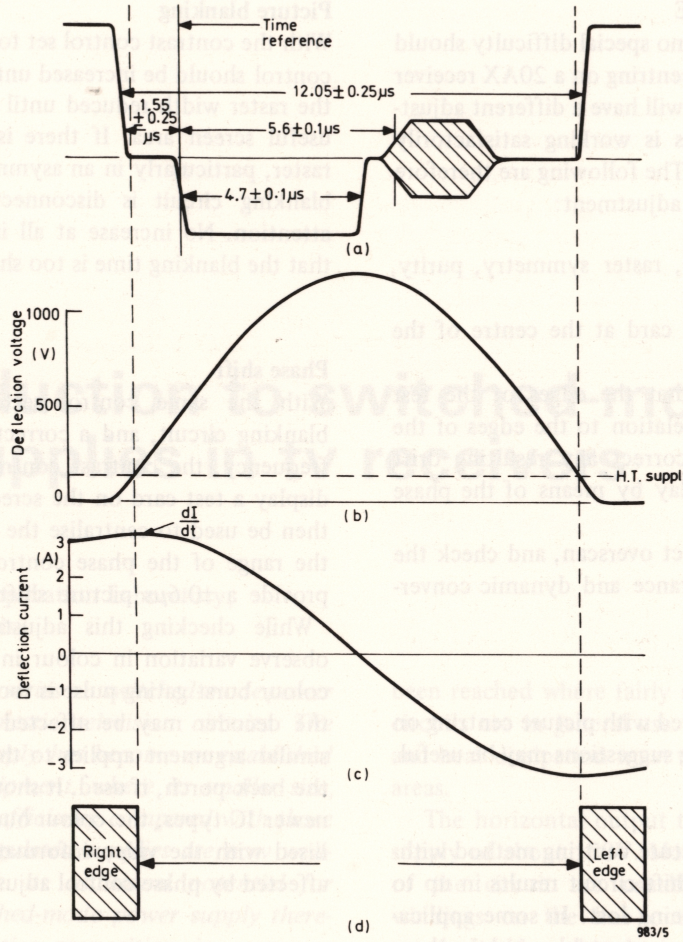


Fig. 5 — Waveforms during the line flyback period:
(a) video (b) scan voltage (c) scan current (d) edges of raster

ADJUSTMENT PROCEDURE

Under normal circumstances no special difficulty should be encountered with picture centring on a 20AX receiver assembly line. Each setmaker will have a different adjustment procedure, and if this is working satisfactorily there is no need to change it. The following are therefore only tentative suggestions for adjustment:

1) adjust static convergence, raster symmetry, purity, and dynamic convergence;
2) set the centre of the test card at the centre of the screen;
3) adjust scan linearity so that the edges of the test card are symmetrical in relation to the edges of the useful screen area, and correct any resulting drift of the centre of the display by means of the phase control;
4) adjust the raster for correct overscan, and check the picture for correct appearance and dynamic convergence.

DIAGNOSTIC TESTS

If any difficulty is encountered with picture centring on 20AX receivers, the following suggestions may be useful.

System capability

As mentioned earlier, the picture centring method without the use of a horizontal shift circuit results in up to 2% of picture information being lost. In some applications such a loss may not be acceptable, and in these cases a picture centring circuit must be added, and an accurate phase control retained.

Static adjustment

The existing prescriptions for purity adjustment should be checked against the latest information available from the set manufacturer, and any recent innovations introduced if necessary.

Picture blanking

With the contrast control set to minimum, the brightness control should be increased until the raster is visible, and the raster width reduced until both edges are within the useful screen area. If there is a significant increase in raster, particularly in an asymmetrical fashion, when the blanking circuit is disconnected, the circuit requires attention. No increase at all in the raster may indicate that the blanking time is too short.

Phase shift

With the same control settings as above, a correct blanking circuit, and a correctly-adjusted line oscillator frequency, the contrast control should be increased to display a test card on the screen. The phase control can then be used to centralise the picture on the screen, and the range of the phase control should be sufficient to provide a ±0.6µs picture shift from the mean position.

While checking this adjustment it is necessary to observe variation in colour information, because if the colour burst gating pulse is too short the performance of the decoder may be affected by the phase control. A similar argument applies to the black level clamping on the back porch, if used. It should be noted that with the newer IC types, the colour burst gating pulse is synchronised with the video information, and therefore is not affected by phase control adjustment.

Range of linearity control

With the centre of the test card at the centre of the screen, it should be possible to obtain a linear display. If the left-hand side of the picture is cramped or extended, and the linearity control is at the limit of its adjustment, it should be checked that the correct type of linearity control is used. This may be a function of the dynamic convergence circuit used.

An introduction to switched-mode power supplies in tv receivers

G.V. SCHAIK,
Philips Semiconductor Application Laboratory,
Nijmegen, The Netherlands.

This article is the first in a series on switched-mode power supplies (SMPS) in domestic television circuits. The switched-mode power supply has become an established part of professional equipment, where its smaller size, lower weight, and higher efficiency compared with those of conventional series-regulated supplies are now well-known advantages. With its additional potential for cost-reduction, the switched-mode power supply therefore becomes an attractive proposition in consumer equipment too, particularly in tv receivers.

The progressive reductions in the size of tv receiver cabinets associated with more compact circuitry and improved picture tube technologies have resulted in heat dissipation becoming a central design consideration. In this context, the switched-mode supply has much to offer.

This first article examines the functioning of the basic SMPS circuits most suitable for use in tv receivers. The feasibility of using these circuits for deriving the 150V supply to the horizontal output stage is discussed, and the possibility of obtaining the horizontal drive itself from the SMPS is also investigated.

INTRODUCTION

The advent of the 110° colour tv picture tube has made it possible to improve the appearance of tv receivers by reducing cabinet depth. Smaller cabinet volumes, particularly those possible with the new 18in (47cm) in-line tubes, impose severe limitations on the size and heat dissipation of receiver circuits. The most power-consuming function in a tv receiver is horizontal deflection. Much design and development effort has been concentrated in this area, and the point has now been reached where fairly standard, optimum-efficiency circuits are in general use. Further reduction in volume and heat dissipation must therefore be obtained in other areas.

The horizontal output transistor requires a stabilised supply of about 150V. Most of the supplies for the rest of the circuit are usually obtained from secondary windings on the line output transformer. The 150V supply itself can be obtained in various ways: such as a mains transformer with a series or parallel regulator, or by thyristor regulation, etc. Unfortunately, all these methods are comparatively inefficient and consequently necessitate considerable heat dissipation. For this reason power supply designers are now turning towards the highly efficient, compact switched-mode power supply (SMPS) for the 150V stabilised supply for the horizontal output stage.

Despite the obvious advantages of the SMPS, and its inherent ability to protect against receiver component failure, most present-day designs fail to exploit its potential to the full. Consequently, many ctv receivers incorporate a more expensive SMPS than is necessary. Optimum design of SMPS circuits for domestic tv is explored in this series of artices.

SWITCHED-MODE POWER SUPPLY PRINCIPLES

An SMPS converts an alternating input voltage into a stabilised d.c. output voltage. The input (mains) voltage is rectified and applied to a switching d.c.-to-d.c. converter (Ref.1). Two types of converter are particularly suitable for application in tv receiver power supplies: the forward converter and the flyback converter.

In both types of converter a choke is used as an energy store. In the forward converter the choke is connected in series with the load: energy is passed to both the choke and the load during the ON period of the converter transistor. In the flyback converter, energy is stored in the choke during the transistor ON period and transferred to the load during the transistor OFF or flyback period. With reference to the choke connection, these converters are sometimes known as series and parallel converters respectively.

Forward converter operation

Fig.1 shows the basic forward-converter circuit, and the associated waveforms and design relationships. When the transistor conducts, the diode is reverse-biased: current flows to the load and the smoothing capacitor through the choke. At the end of the conduction period, the energy in the choke causes the current flow to continue via the diode. Because of this action, the diode is described as a flywheel diode. Since the current level through the choke at the end of the conduction period, and hence the energy stored in it, is a function of the transistor ON time, the rate of transfer of energy to the load, and hence the output voltage, can be controlled by varying the ON time of the transistor at a given frequency.

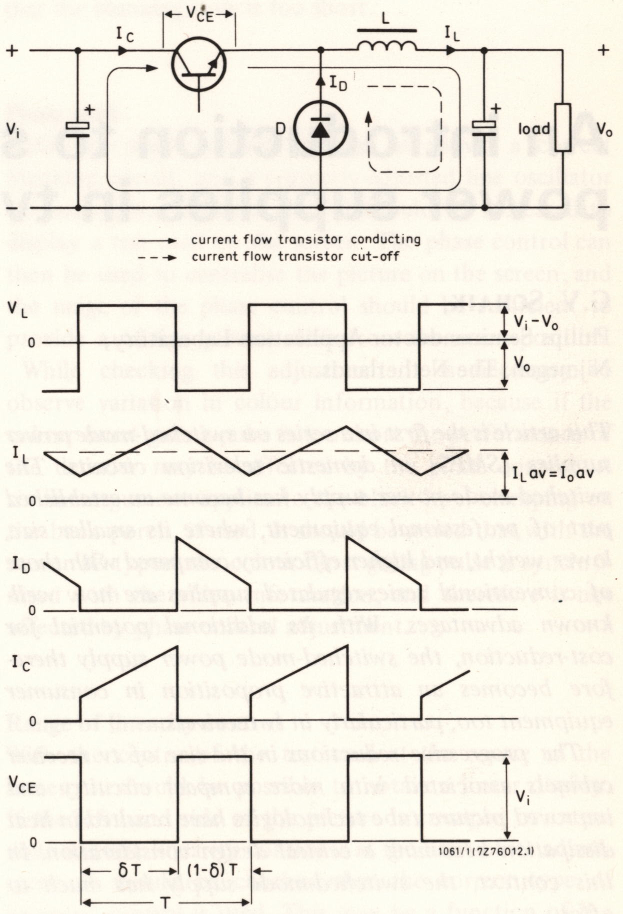
When the flywheel diode is conducting, the voltage across the choke is roughly equal to the output voltage, and low-level stabilised supplies can be obtained by adding secondary windings to the choke with rectifier diodes. (Although a choke with secondary windings has the appearance of a transformer, its function is still best considered as that of a choke and it is referred to as such in these articles.) It should be noted that when secondary windings are used, the amount of energy available for transfer to the main output per cycle is reduced.

To minimise the internal impedance of the converter, the choke current should continue to flow throughout the transistor OFF period. The minimum inductance for the choke is the value that restricts the peak-to-peak value of the choke ripple current to twice the minimum required average output current.

Mains isolation of this type of circuit requires the addition of a power transformer and rectifier diode but this is usually not an economically attractive proposition in tv receivers.

Flyback converter operation

Fig.2 gives the basic circuit diagram of the flyback converter together with the associated waveforms and design relationships. When the transistor conducts, the supply voltage is connected across the choke and the diode is reverse-biased. The choke current increases during transistor ON period, storing energy. When the transistor turns off, the voltage across the choke reverses and the diode becomes forward-biased. Current then flows to the output capacitor and the load. By varying the conduction time of the transistor at a given switching frequency, the amount of energy stored in the choke during each ON period, and hence the average level of the output voltage, can be controlled.



$$V_o = \delta V_i.$$

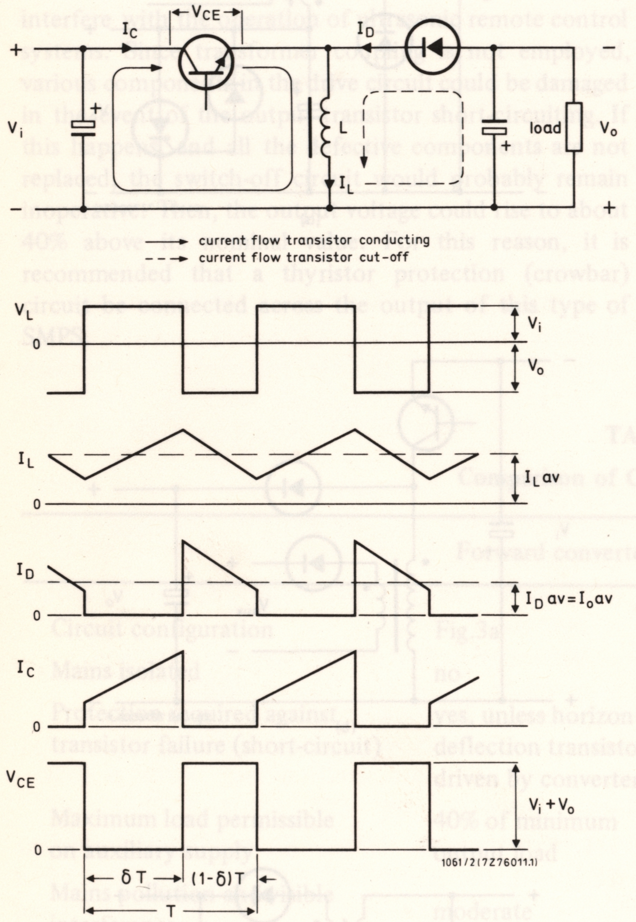
$$L_{min} = \frac{V_o{}^2 T}{2P_{o\,min}} \left(1 - \frac{V_o}{V_{i\,max}} \right).^*$$

$$I_{C\,peak} = \frac{V_o T}{2L} \left(1 - \frac{V_o}{V_{i\,max}} \right) + \frac{P_{o\,max}}{V_o}.$$

$$V_{CE\,max} = V_{i\,max}.$$

*For continuous choke current.

Fig.1 — Basic circuit of a forward d.c.-to-d.c. converter with associated waveforms and design relationships



$$V_o = \frac{\delta V_i}{1 - \delta}.$$

$$L_{min} = \left(\frac{V_{i\,max} V_o}{V_{i\,max} + V_o} \right)^2 \times \frac{T}{2P_{o\,min}}.^*$$

$$I_{C\,peak} = P_{o\,max} \left(\frac{V_{i\,min} + V_o}{V_{i\,min} V_o} \right) + \frac{T}{2L} \left(\frac{V_{i\,min} V_o}{V_{i\,min} + V_o} \right).$$

$$V_{CE\,max} = V_{i\,max} + V_o.$$

*For continuous choke current.

Fig.2 — Basic circuit of a flyback d.c.-to-d.c. converter with associated waveforms and design relationships

In the flyback converter, all the energy to be transferred to the output capacitor and load is first stored in the choke. It is therefore possible to achieve mains isolation by adding a secondary winding to the choke. If this is done, ringing of the transistor collector-emitter voltage must be minimised by reducing the leakage inductance between the choke windings. Tight coupling of the windings will also improve the regulation of the circuit.

To minimise the internal impedance of the converter, the choke inductance must be high enough to ensure that the diode conducts throughout the transistor OFF period.

One disadvantage of this type of converter is that the output capacitor is charged only during the transistor OFF period. Thus the output capacitor ripple current is much greater than with the forward converter.

CHOICE OF CONVERTER

At this point we have a choice between three types of converter. These are shown in Fig.3. Inevitably each has advantages and disadvantages, and the final choice will be a compromise between cost-effectiveness and other receiver design requirements. The three types of converter are compared in more detail below and in Fig.3.

Performance of forward converter

In the forward converter (Fig.3a) part of the output power is supplied directly to the load during the transistor ON period. Moreover, the peak value of V_{CE} that the output transistor must withstand is equal to the d.c. supply voltage. These factors impose lower demands on the transistor and diode in a forward converter compared with a flyback converter. Furthermore, the required choke core cross-section is smaller and the capacitor ripple current lower. Also because the voltage being switched is lower, mains interference is slight.

The disadvantages of the forward converter for tv applications are as follows. Firstly, mains isolation cannot be achieved simply by adding a secondary winding to the choke. Secondly, if the switching transistor becomes internally short-circuited, the full input voltage is applied direct to the load. For this latter reason, unless the horizontal output transistor takes its drive from the SMPS, a thyristor 'crowbar' circuit must be connected across the output as shown in Fig.3a. If the output voltage then exceeds a certain level, the thyristor conducts and causes the mains input fuse to rupture. A third disadvantage is the fact that the load on the auxiliary supply is limited to about 40% of the minimum load on the main output. Thus, if the main load is disconnected as a result of a fault, the auxiliary supply also becomes unavailable.

Performance of flyback converter

Figs.3b and 3c show two possible configurations for the flyback converter. Note that in the circuit of Fig.3b, the transistor could be connected to a tap on the primary winding of the choke. The collector voltage of the transistor would then need to be clamped against overswing.

The advantages of the flyback converter are the opposite of the disadvantages of the forward converter. Mains isolation can be achieved simply by adding a secondary winding to the choke. It is not necessary to protect the load against excess voltage in the event of a transistor short-circuit. The load permissible on the auxiliary supply is not related to the main output load. If the main load is disconnected, the auxiliary supply is still available; this facilitates fault diagnosis.

Similarly, the disadvantages of the flyback converter complement the advantages of the forward converter. The peak value of V_{CE} that the transistor must withstand is the sum of the input and output voltages, and the choke has to store all the energy delivered during one ON period of the transistor. Thus the transistor and diode both have to pass a higher peak current and withstand a higher peak voltage. Furthermore, a larger cross-section core is required for the choke, and the output capacitor has to pass a higher ripple current. Because higher currents and voltages are being switched, the circuit generates more mains interference.

Comparison of performance

Table 1 summarises the main features of the three circuits given in Fig.3. The information presented in Table 1 is based on the following power supply specification. (The output power includes all output circuit losses.)

Mains voltage	190 to 260V r.m.s.
D.C. input	240 to 350V
Output voltage	150V d.c.
Output power*	100W min. 175W max.
Switching period	64μs (f = 15.62kHz)

SMPS DRIVE CIRCUITS

The need for silent operation sets the lower limit of the operating frequency of the SMPS switch at about 16kHz. The upper limit is set at about 50kHz by the active switching components and the choke losses. The type of drive circuit used depends on whether the operating frequency is to be constant or to vary in response to mains voltage or load fluctuations.

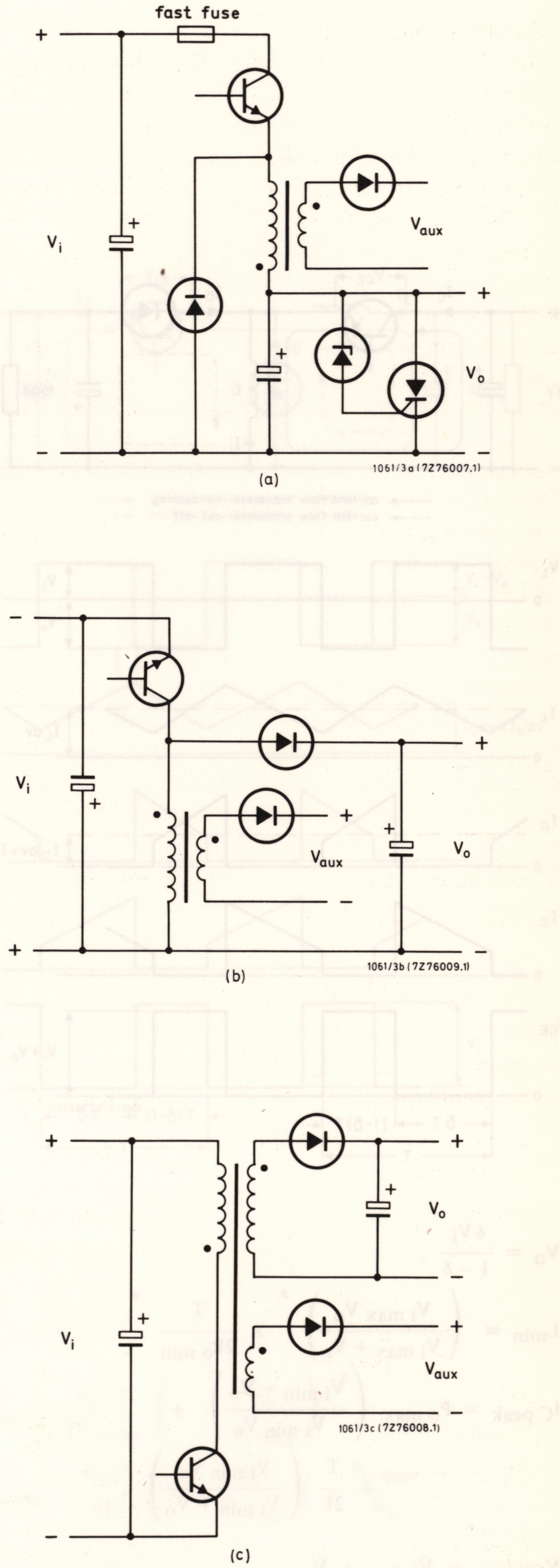


Fig.3 — Choice of basic d.c.-to-d.c. converter. (a) Forward converter with protection against output transistor short circuit. (b) Non-isolated flyback converter with auxiliary supply. (c) Fully isolated flyback converter

---

*Including output-circuit losses.

Variable-frequency drive

The simplest form of SMPS is the self-oscillating type (Ref.2). In such a circuit, the output transistor is first turned on by a starting pulse and then driven from a winding on the output choke. When its collector current exceeds a certain level, the output transistor is turned off.

The switching frequency of this type of converter varies with mains voltage and load changes, and may interfere with the operation of ultrasonic remote control systems. Since transformer coupling is not employed, various components in the drive circuit could be damaged in the event of the output transistor short-circuiting. If this happens, and all the defective components are not replaced, the switch-off circuit would probably remain inoperative. Then, the output voltage could rise to about 40% above its nominal value. For this reason, it is recommended that a thyristor protection (crowbar) circuit be connected across the output of this type of SMPS.

In a self-oscillating SMPS, the drive pulses will necessarily be inhibited if the output transistor fails. This so complicates fault-finding that replacement of the entire circuit is usually necessary. Thus, modular construction is recommended.

Fixed-frequency drive

To operate at a constant frequency, an SMPS must incorporate a separate driver transistor and an oscillator. The latter is usually an integrated circuit. The driver transistor can be coupled to the output transistor either directly or through a transformer. Transformer coupling is generally preferred since it improves circuit efficiency and reliability.

In principle, the operating frequency may be freely chosen within wide limits: however, in tv receiver applications, it is recommended that the SMPS drive be synchronised with the horizontal flyback pulses. This is discussed later.

TABLE 1

Comparison of Converter Parameters

	Forward converter	Flyback converter	
Circuit configuration	Fig.3a	Fig.3b	Fig.3c
Mains isolated	no	no	yes
Protection required against transistor failure (short-circuit)	yes, unless horizontal deflection transistor driven by converter	not required	
Maximum load permissible on auxiliary supply	40% of minimum output load	not related to output load	
Mains pollution and visible interference	moderate	acceptable	serious
Serviceability	without main load no auxiliary supply	auxiliary supply available without main load	
Converter switching transistor peak $V_{CE\,max}$	350V	500V	575V
peak $I_{C\,max}$	1.6 to 1.2A*	2.75 to 1.9A*	2.1A
Duty factor range to maintain constant output	0.43 to 0.63	0.30 to 0.39	0.39 to 0.48
Minimum choke inductance	4.1mH (without auxiliary load)	3.5mH	6mH
Core of output choke	U30 (Ferroxcube 3C8) or iron	U30 (Ferroxcube 3C8) or iron	E55 (Ferroxcube 3C8)
Output ripple current	low	high	high

*In this circuit arrangement the maximum value of the inductance is not limited. The second figure is the minimum I_C that can be achieved.

Overload protection

If too much current is drawn from an SMPS, the output transistor may be damaged. Some form of protection against overload should be incorporated in the drive circuit. This could be afforded by a simple arrangement to reduce the duty factor of the output transistor so that its average collector current remains at a normal level during an overload (Ref.3). With such an arrangement, however, the peak value of the collector current during a load short-circuit depends on the impedance of the collector circuit and the minimum duty factor that can be achieved. The minimum value of duty factor is set by the minimum duty factor of the control circuit and the storage time of the output transistor. In consequence, dissipation continues during a short-circuit and the risk of damage to the output transistor remains.

To overcome this problem, a special circuit is used to switch off the power supply completely when its output current exceeds a preset level (Ref.4). After a short delay, this circuit switches the supply back on again, but with a greatly reduced duty factor (inrush limiting) that gradually increases. If the overload persists, the supply is again turned off and the cycle repeats. Because of its mode of operation, this arrangement has become known as a 'hiccup' circuit. Some of the waveforms associated with it are given in Fig.4.

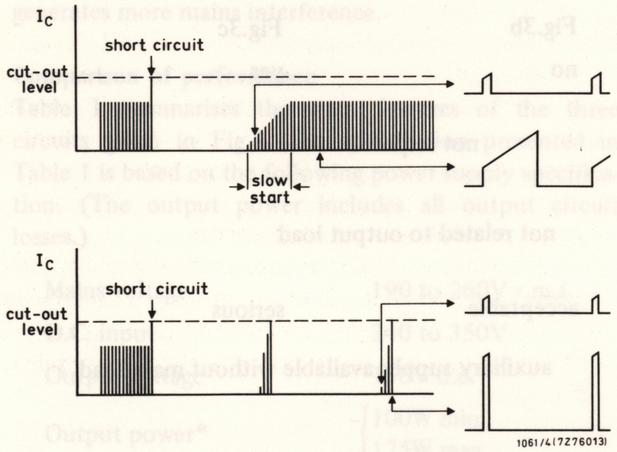


Fig.4 – Operation of the 'hiccup' circuit for overload protection. On the left is the SMPS output transistor collector current. This is shown with an expanded time scale on the right. Conditions are (a) transient short circuit; and (b) persistent short circuit

If low-voltage supplies are also derived from the SMPS, the currents drawn from them would have to increase to several times their nominal value before the main overload protection circuit would operate. Low-voltage supply circuits must, therefore, include fuses which will not only protect the rectifier diodes and transformer windings, but will also give a positive indication of the location of any fault.

Drive circuits

For reliable operation of a tv receiver with an SMPS, the drive circuit must provide the correct base drive for the output transistor under all operating conditions. Furthermore, there must be:

- provision for synchronising the SMPS drive pulses with the horizontal oscillator;
- control of the duty factor of the SMPS output transistor to stabilise the supply to the horizontal deflection circuit;
- an error amplifier for monitoring the gain and stability of the feedback loop;
- inrush-current limiting at switch-on;
- a reference voltage for comparison with the feedback voltage;
- limitation of the maximum duty factor of the drive pulses.

A circuit that meets all these requirements is, inevitably, rather complex and would require the use of a large number of discrete components, particularly in the case of a fixed-frequency SMPS. For this reason, the following SMPS driver ICs have been developed.

TDA2640

The TDA2640 is a control IC for fixed-frequency SMPS circuits. It includes:

- a fixed-frequency oscillator with synchronising facility;
- duty-factor control;
- 'slow-start' to limit inrush current at switch-on;
- overcurrent protection (hiccup circuit);
- protection against low supply voltage;
- overvoltage protection for no-load conditions.

TDA2581

This SMPS control IC has the same facilities as the TDA2640, but provision is made for phase control of the oscillator. This allows the drive for the horizontal deflection transistor to be obtained from a secondary winding on the output choke of the SMPS.

INTERFERENCE

Interference considerations fall into two main categories. These are interference that the SMPS may inject into the mains, and interference caused by the SMPS that affects the picture on the tv receiver itself.

Mains pollution

A certain amount of interference may be injected into the mains supply as a result of the high switching frequency and power operating level of an SMPS (Ref.5). This interference falls into two categories, symmetrical and asymmetrical.

Symmetrical interference appears between the mains input lines. It is caused by ripple current flowing through the mains rectifier section of the SMPS.

Symmetrical interference can be kept to 15 or 20dB below the maximum level permitted by the CISPR (Comité International Spécial des Perturbations Radioélectriques) by including a simple LC filter between the mains supply and the input to the SMPS. The filter should be located as close as possible to the SMPS input.

Asymmetrical mains pollution appears between earth and each of the mains lines; this type of interference is caused by current flowing through the capacitances associated with components carrying high alternating voltages. It is most pronounced in mains-isolated power supplies because of the capacitance between the primary and secondary windings of the output choke.

To reduce asymmetrical interference, the capacitances between alternating voltage carrying components and earth must be reduced. In a mains-isolated supply, there is a high alternating voltage between the (isolated) chassis and earth, because of the interwinding capacitance. The capacitance between chassis and earth is fairly high but the interference can be reduced by connecting a capacitor between the chassis and the primary (live) side of the power supply. However, because of the mains-isolation requirements the value of this capacitor is limited to about 5nF. It is essential, therefore, to minimise the interwinding capacitance itself. This can be achieved by placing two screens between the windings: one screen connected to the 'cold' end of the primary, and the other to the chassis.

Visible interference

Even when a tv receiver is not equipped with an SMPS, one of the largest problems is pick-up of spurious signals (such as spikes from the horizontal deflection) by the tuner and i.f. stages. This type of interference is particularly severe when the set is tuned to a weak signal at the low-frequency end of the v.h.f. band.

In a receiver with an SMPS, the main cause of this type of interference is the fast rise and fall of the current pulses passed by the various switching devices. Visible interference due to spurious pick-up can be minimised by:

- keeping the length of the switching loop small;
- connecting h.f. decoupling capacitors across electrolytic capacitors;
- using soft-recovery diodes;
- using non-inductive resistors.

SWITCHED-MODE POWER SUPPLY FOR HORIZONTAL DEFLECTION CIRCUIT

An SMPS is, in principle, a self-contained unit. However, problems can arise when the load varies periodically, as with a horizontal output stage.

Since the load then varies at line frequency, the SMPS output itself may be modulated by way of the control loop. This results in beat frequencies between the SMPS switching frequency and the line frequency (and between their harmonics and sub-harmonics) being superimposed on the SMPS output. If the SMPS is not synchronised with the horizontal oscillator, an LC filter must be connected to its output.

If careful consideration is not given to the layout and screening of an SMPS, the receiver tuner and i.f. stages will pick up interference spikes. Where the SMPS is synchronised with the horizontal oscillator, this interference can appear as a single vertical bar. Where the SMPS is not synchronised, a number of bars or dots can appear on the screen. These will generally be less obvious than the single bar. However, with asynchronous operation it is essential to ensure that higher-order harmonics of the SMPS switching and line frequencies remain unrelated.

It is clear therefore that the SMPS switching frequency should be synchronised with the horizontal oscillator. This has two other main advantages:

- a simpler smoothing circuit can be used at the output of the SMPS, and
- the smaller value smoothing capacitor stores less energy and therefore improves flashover behaviour.

Where mains isolation is required, the synchronisation signal can be obtained from a single-turn secondary on the output choke.

Typical non-isolated forward converter

A simplified circuit diagram of a non-isolated forward-converter SMPS is shown in Fig.5. The drive to the converter transistor is provided by the TDA2640 through a driver transistor. A TDA2581 could also be used. A crowbar circuit is provided for load protection.

Typical non-isolated flyback converter

One of the many possible arrangements of a non-isolated flyback-converter SMPS is shown in Fig.6. Here, the use of a separate power supply for the drive circuit is avoided by connecting the receiver chassis to the positive side of the input.

Typical mains-isolated flyback converter

Of the possible circuit arrangements for a mains-isolated flyback SMPS, that chosen depends largely on the method used to obtain the error feedback signal. The circuit of Fig.7 is arranged to eliminate the separate supply for the driver stage: the error signal is derived from a separate winding on the output choke.

DRIVING HORIZONTAL DEFLECTION FROM SMPS

We have already seen that there are considerable advantages to be gained from synchronising the SMPS converter with the horizontal deflection stage. A further component saving can be realised by taking the next logical step: that is, using the SMPS drive circuit to drive

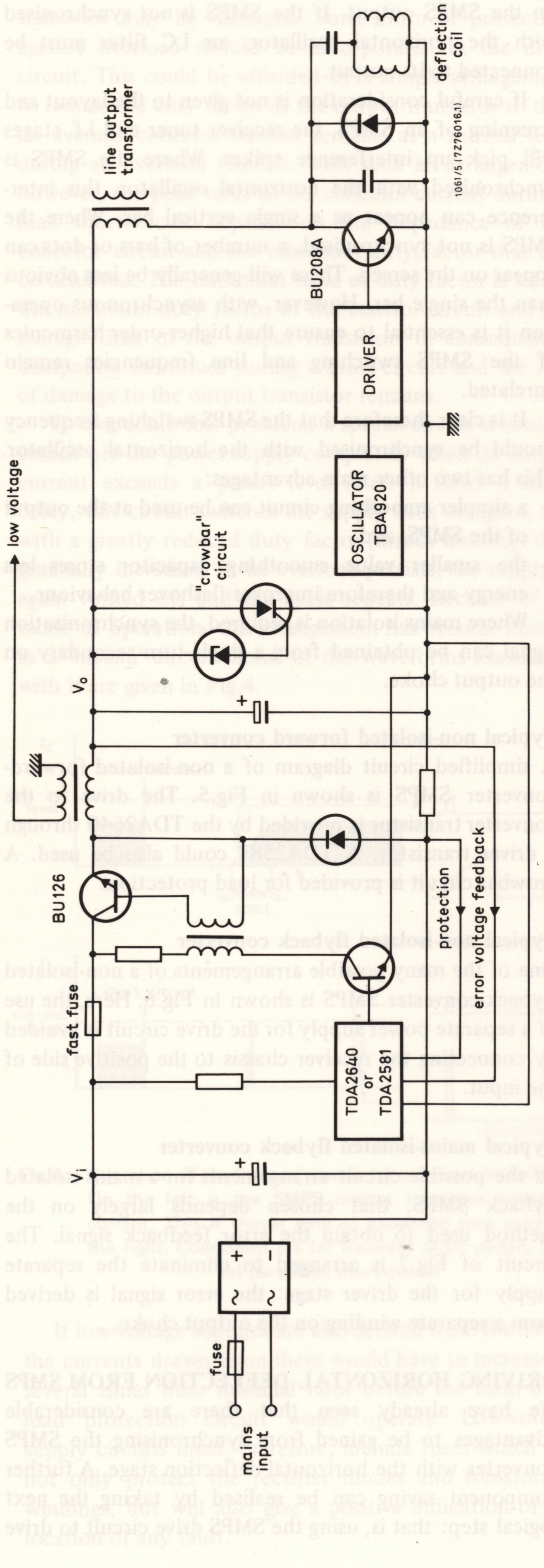

low voltage
V_o
'crowbar' circuit
BU126
fast fuse
mains input
fuse
TDA2640 or TDA2581
protection
error voltage feedback
OSCILLATOR TBA920
DRIVER
BU208A
line output transformer
deflection coil
1081/5 (7Z76016)

Fig.5 – Simplified circuit diagram of an SMPS using a forward converter for supplying power to a horizontal deflection circuit

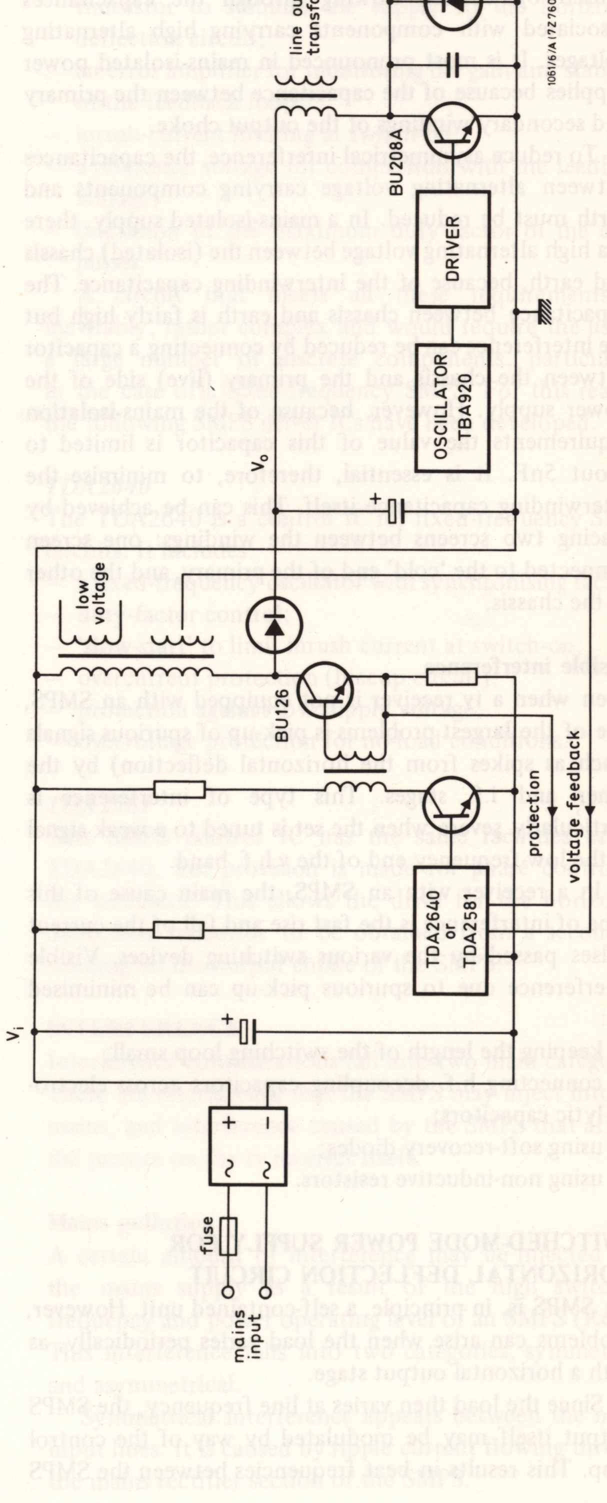

low voltage
V_o
V_i
BU126
fuse
mains input
TDA2640 or DA2581
protection
voltage feedback
OSCILLATOR TBA920
DRIVER
BU208A
line output transformer
deflection coil
1081/6A (7Z76017)

Fig.6 – Simplified circuit diagram of an SMPS using a flyback converter for supplying power to a horizontal deflection circuit

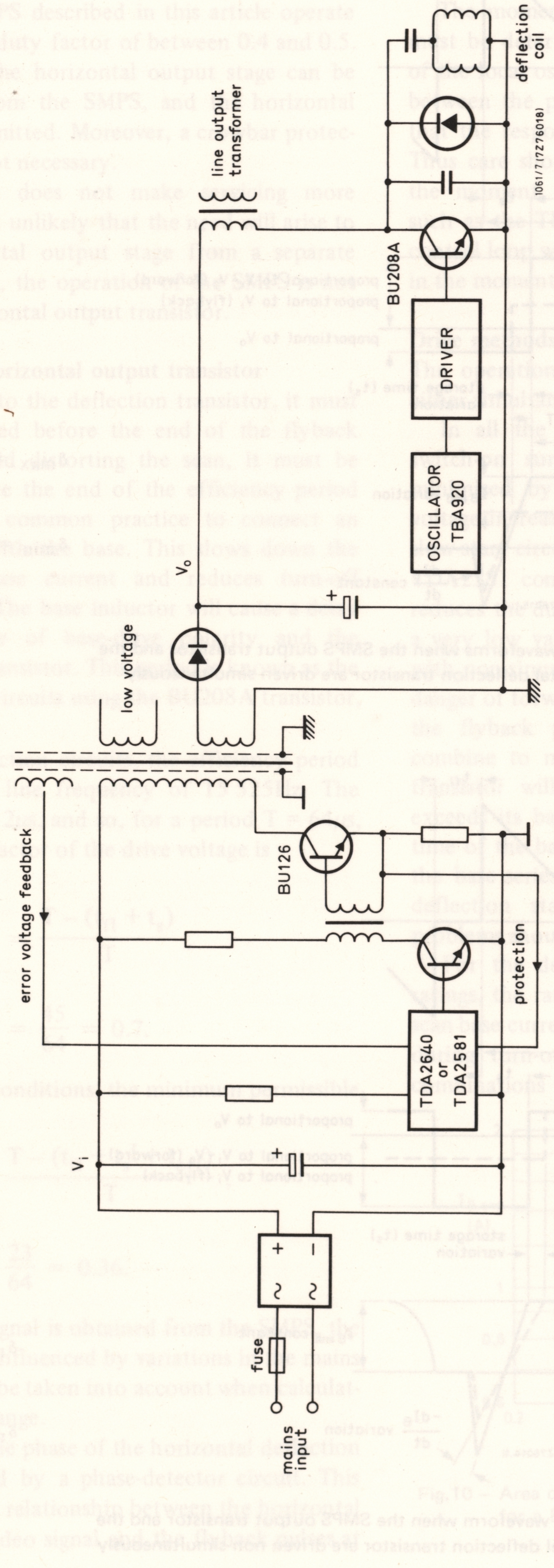

low voltage
V_o
V_i
error voltage feedback
BU126
fuse
mains input
TDA2640 or TDA2581
protection
OSCILLATOR TBA920
DRIVER
BU208A
line output transformer
deflection coil
1061/7 (7Z76018)

Fig.7 – Simplified circuit diagram of a mains-isolated SMPS using a flyback converter for supplying power to a horizontal deflection circuit

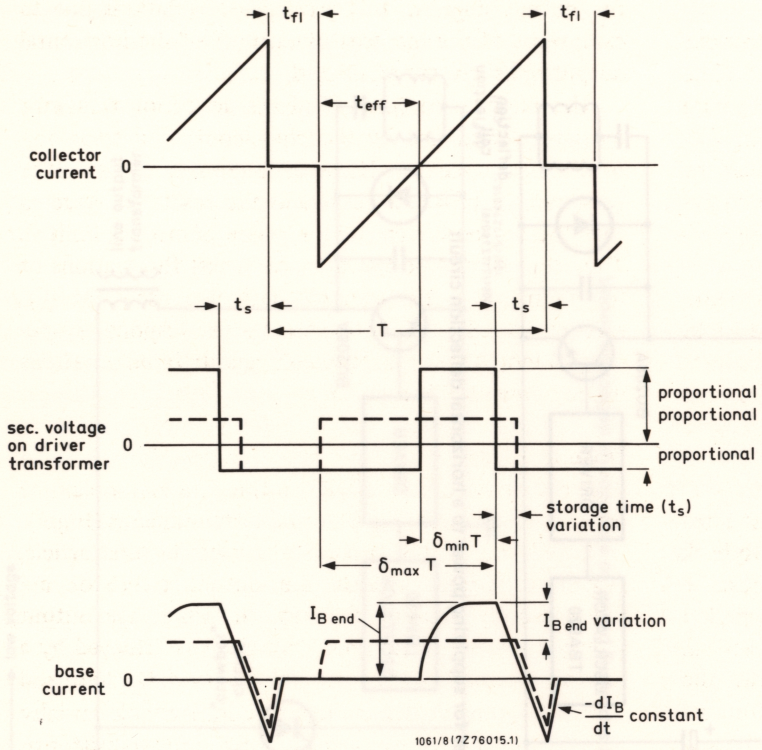

t_{fl}
t_{fl}
t_{eff}
collector current
0
t_s
t_s
T
sec. voltage on driver transformer
0
proportional to $V_i - V_o$ (forward)
proportional to V_i (flyback)
proportional to V_o
storage time (t_s) variation
$\delta_{min}T$
$\delta_{max}T$
$I_{B\,end}$
$I_{B\,end}$ variation
base current
0
$\frac{-dI_B}{dt}$ constant
1061/8(7Z76015.1)

Fig.8 — Base drive waveforms when the SMPS output transistor and the horizontal deflection transistor are driven simultaneously

$$\delta_{max} = \frac{T - (t_{fl} + t_s)}{T}\,.$$

$$\delta_{min} = \frac{T - (t_{fl} + t_s + t_{eff})}{T}$$

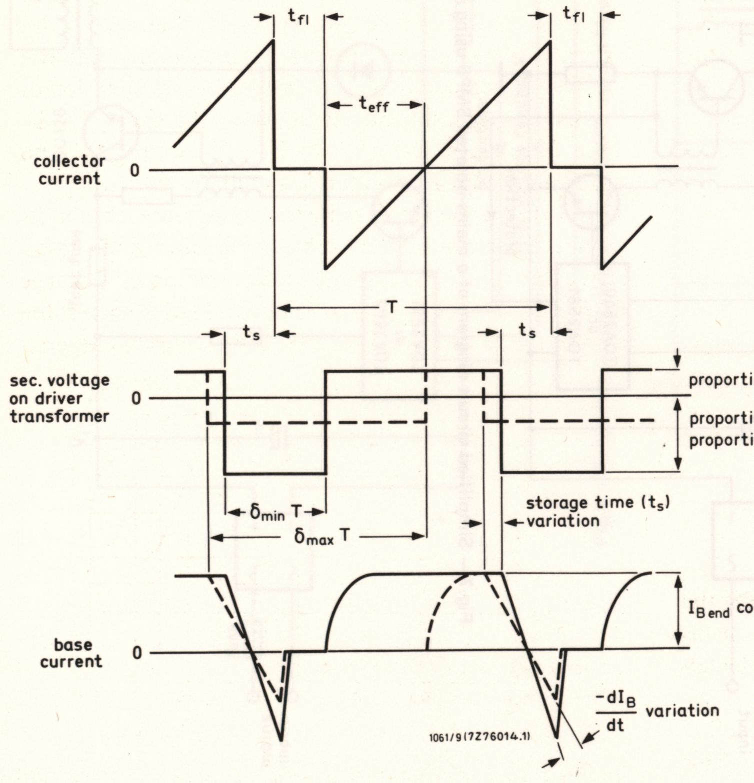

t_{fl}
t_{fl}
t_{eff}
collector current
0
T
t_s
t_s
sec. voltage on driver transformer
0
proportional to V_o
proportional to $V_i - V_o$ (forward)
proportional to V_i (flyback)
storage time (t_s) variation
$\delta_{min}T$
$\delta_{max}T$
$I_{B\,end}$ constant
base current
0
$\frac{-dI_B}{dt}$ variation
1061/9(7Z76014.1)

Fig.9 — Base drive waveform when the SMPS output transistor and the horizontal deflection transistor are driven non-simultaneously

$$\delta_{max} = \frac{t_s + t_{fl} + t_{eff}}{T}$$

$$\delta_{min} = \frac{t_s + t_{fl}}{T}$$

the horizontal deflection stage as well.

The base drive waveform for a horizontal output transistor must have a duty factor of about 0.5. The various types of SMPS described in this article operate satisfactorily with a duty factor of between 0.4 and 0.5. Thus the drive to the horizontal output stage can be obtained directly from the SMPS, and the horizontal drive stage can be omitted. Moreover, a crowbar protection circuit is then not necessary.

This arrangement does not make servicing more difficult because it is unlikely that the need will arise to operate the horizontal output stage from a separate supply. Furthermore, the operation of the SMPS is not affected by the horizontal output transistor.

Base drive for the horizontal output transistor

To prevent damage to the deflection transistor, it must not be forward-biased before the end of the flyback period, and, to avoid distorting the scan, it must be forward-biased before the end of the efficiency period t_{eff} (Ref.6). It is common practice to connect an inductor in series with the base. This slows down the decrease of the base current and reduces turn-off dissipation (Ref.7). The base inductor will cause a delay between the change of base-drive polarity and the turning off of the transistor. This period is known as the storage time, t_s. In circuits using the BU208A transistor, $t_s \simeq 7\mu s$ (Ref.8).

In practical deflection circuits, the efficiency period $t_{eff} \simeq 22\mu s$ for a line frequency of 15 525Hz. The flyback time $t_{fl} = 12\mu s$, and so, for a period $T = 64\mu s$, the maximum duty factor of the drive voltage is

$$\delta_{max} = \frac{T - (t_{fl} + t_s)}{T},$$

$$= \frac{45}{64} = 0.7.$$

Under the same conditions, the minimum permissible duty factor is

$$\delta_{min} = \frac{T - (t_{fl} + t_s' + t_{eff})}{T},$$

$$= \frac{23}{64} = 0.36.$$

When the drive signal is obtained from the SMPS, the storage time (t_s') is influenced by variations in the mains voltage. This should be taken into account when calculating the duty factor range.

In a tv receiver, the phase of the horizontal deflection current is controlled by a phase-detector circuit. This determines the phase relationship between the horizontal sync pulses in the video signal and the flyback pulses at the output stage. In this way, phase variations due to excursions of the moment of turn-off of the horizontal output transistor are minimised.

The moment of turn-off of the deflection transistor must be determined by the phase-controlled transition of the local oscillator. However, there is a lowpass filter between the phase detector and the reactance stage so that the response time of the phase control is limited. Thus care should be taken to avoid fast fluctuations of the moment of turn-off. (Modern line oscillator ICs, such as the TBA920, incorporate a fast-response phase-control loop which is able to compensate rapid variations in the moment of turn-off.)

Drive methods

The operation of the SMPS output transistor can be either simultaneous (Fig.8), or non-simultaneous (Fig.9).

In all the SMPS circuits discussed in this article, switch-on surges through the output transistor are minimised by limiting the rate at which the output voltage increases to its final value. This is achieved by a slow-start circuit of the type found in the TDA2640 and TDA2581 control ICs. As the slow-start circuit initially reduces the duty factor of the SMPS output transistor to a very low value and allows it to increase only slowly, with non-simultaneous drive there would appear to be a danger of forward-biasing the deflection transistor during the flyback period (Fig.9). In practice, two factors combine to make this unlikely: firstly, the deflection transistor will conduct only when the applied drive exceeds its base-emitter voltage; and secondly, the rise time of the base current is increased by the inclusion of the base series inductor. Moreover, as the energy in the deflection stage during switch-on is quite low, no problems should arise with non-simultaneous drive.

For the deflection transistor to operate within its ratings, the range of permissible combinations of end-of-scan base current ($I_{B\,end}$) and base-current rate of change during turn-off (dI_B/dt) is limited. Fig.10 shows the combinations permitted for a BU208A transistor operat-

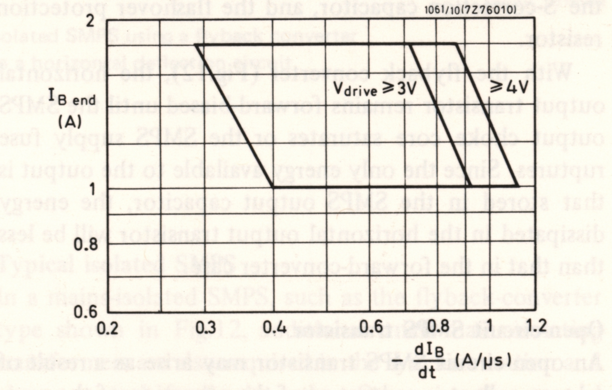


Fig.10 — Area of permissible combination of $I_{B\,end}$ and dI_B/dt for a BU208A transistor operated with a peak collector current of 4A

ing with a peak collector current of 4A. Both parameters may be influenced by mains-voltage fluctuations, depending on the type of drive circuit used.

In simultaneous operation, Fig.8, the amplitude of the positive drive voltage depends on the level of the incoming mains, and so, therefore, does the value of $I_{B\ end}$. However, the amplitude of the negative drive voltage depends on the stabilised output of the SMPS, and thus dI_B/dt during turn-off is independent of the mains voltage.

In non-simultaneous operation, the situation is reversed. The value of $I_{B\ end}$ is now determined by the level of the SMPS output, and is independent of mains voltage. However, the amplitude of the negative drive voltage is now mains-dependent, and so is the value of the turn-off dI_B/dt.

Operation during fault conditions

Where a horizontal output stage is both powered and driven by an SMPS, there are four important fault conditions: these are simultaneous drive with SMPS transistor either open-circuit or short-circuit, and non-simultaneous drive with SMPS transistor either open-circuit or short-circuit. Since the horizontal output transistor itself cannot be damaged by a fault that causes it to be cut off by reverse base drive, there are only two of these fault conditions which could cause subsequent more extensive damage. These are simultaneous drive with a short-circuited SMPS transistor, and non-simultaneous drive with an open-circuit SMPS transistor.

Short-circuited SMPS transistor

The short-circuited SMPS transistor is the most common fault condition. Its effects on circuits with simultaneous drive are as follows.

With the forward converter (Fig.11), the horizontal output transistor remains forward-biased until the SMPS output voltage has risen to the level of the rectified mains voltage. During this time, the level of the peak collector current of the horizontal output transistor is determined by the values of the SMPS output capacitor, the S-correction capacitor, and the flashover protection resistor.

With the flyback converter (Fig.12), the horizontal output transistor remains forward-biased until the SMPS output choke core saturates or the SMPS supply fuse ruptures. Since the only energy available to the output is that stored in the SMPS output capacitor, the energy dissipated in the horizontal output transistor will be less than that in the forward-converter case.

Open-circuit SMPS transistor

An open-circuit SMPS transistor may arise as a result of a loose collector contact or of the operation of the overload protection circuit. (It is usual in remote-controlled tv receivers to use the protection circuit as an on/off switch.) The effect of this on circuits with a non-simultaneous drive is as follows.

With forward and flyback converters, the horizontal output transistor will remain forward-biased until all the energy has been removed from the SMPS output choke. A choke of minimum inductance (L_{min}) will minimise the dissipation in the horizontal output transistor. It is possible that the deflection transistor could be forward-biased during the flyback period when its collector voltage is very high. This condition can be prevented by applying a negative-going pulse derived from the flyback pulse to the base of the SMPS driver transistor. However, this will not prevent the deflection transistor from being forward-biased during its storage time. If this does happen, the deflection transistor could be switched off fast at a high collector current as a result of feedback. (Usually the collector current is limited by the gradual increase of collector-emitter saturation voltage so that the flyback voltage is heavily damped.) Since fast switch-off would result in a high collector voltage peak, it is apparent that a carefully designed protection circuit must be used.

With a forward converter, forward-biasing of the deflection transistor during the flyback interval can be prevented if provision is made for the protection circuit to be operated by the flyback pulse. A voltage for this purpose can be developed across a resistor connected in series with the flywheel diode, but this method has several disadvantages. Firstly, an overcurrent cannot be detected when the SMPS output transistor is conducting: a low-inductance (L_{min}) choke may therefore allow the SMPS output transistor to pass excess current before the protection circuit can operate. Secondly, other inputs (such as overvoltage) will not operate the protection circuit.

The foregoing observations regarding the supply and driving of a horizontal output stage from an SMPS are summarised in Table 2. In compiling this table it has been assumed that the drive voltage amplitude at nominal mains input is at least 3V. Furthermore, the amplitude of the negative drive voltage has been taken to be less than 5V and therefore the emitter-base current remains below the maximum permitted average value of 100mA.

Typical non-isolated SMPS

Fig.11 gives the basic elements of a non-isolated forward-converter SMPS for the supply of both power and drive to a horizontal deflection circuit. The drive for the SMPS transistor is obtained from the TDA2581, described earlier. The phase detector in this IC is used to derive a control signal for the horizontal oscillator from the horizontal flyback pulses and the sync pulses from horizontal combination TDA2571. The control loop so formed has a short time-constant, allowing it to be used to minimise the effect of horizontal output transistor storage time variations on the phase of the deflection current. This is done by using the output of the phase detector to control the horizontal oscillator which, in turn, controls the turn-off instant of the horizontal output transistor. The control circuit of the SMPS varies the turn-off instant of the SMPS output transistor and, thus, the turn-on instant of the horizontal output transistor. For this reason, the TDA2581 is suitable only for circuits where complementary drive is used for the horizontal output transistor.

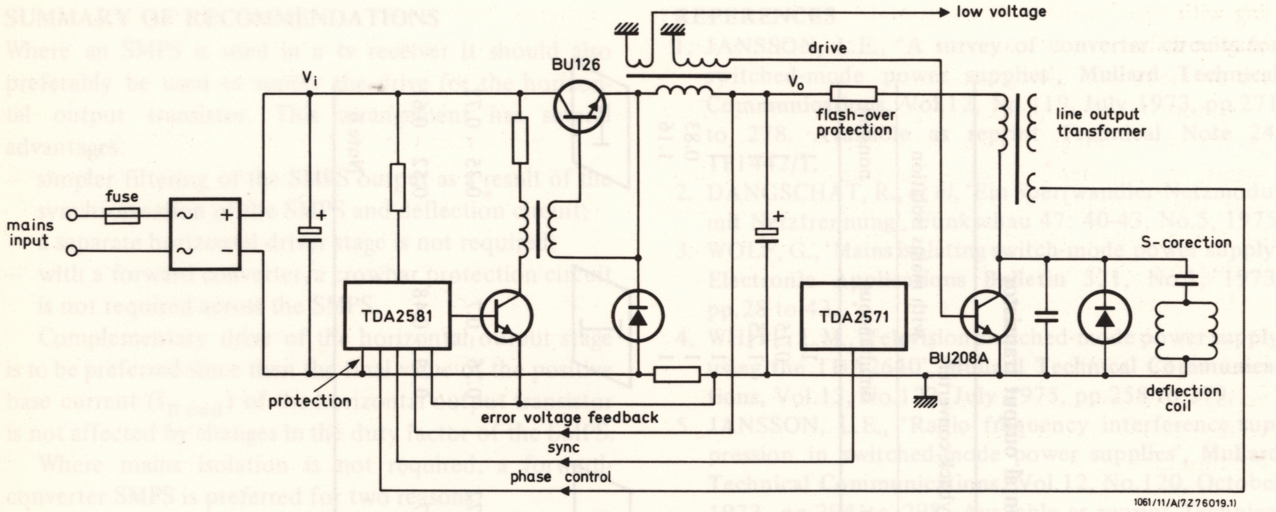


Fig.11 — Simplified circuit diagram of an SMPS using a forward converter to supply power and drive to a horizontal deflection circuit

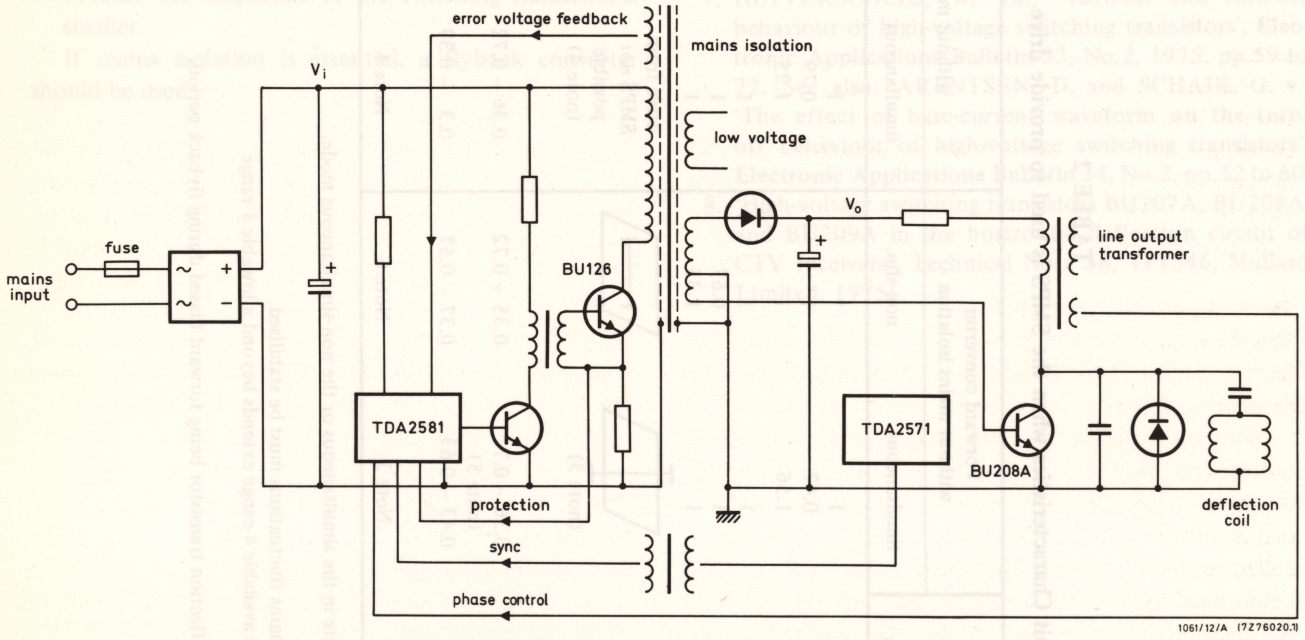


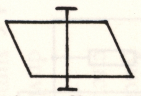
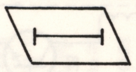
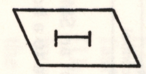
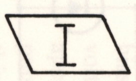
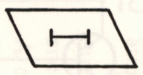
Fig.12 — Simplified circuit diagram of a mains-isolated SMPS using a flyback converter for supplying power and drive to a horizontal deflection circuit

Typical isolated SMPS

In a mains-isolated SMPS, such as the flyback-converter type shown in Fig.12, additional small mains-isolating transformers are also required in the synchronisation and phase control feedback paths. Otherwise the control arrangement using the TDA2581 control IC is similar to that used in the previous example.

TABLE 2

Summary of Circuit Characteristics when the SMPS is used to provide drive for the horizontal output transistor

		Forward converter without mains isolation		Flyback converter without mains isolation		Flyback converter with mains isolation	
Type of drive (note 1)	V_{mains}	simultaneous	non-sim.	simultaneous	non-sim.	simultaneous	non-sim.
Base current I_B end (normalised)	nom.	1	1	1	1	1	1
	190V	0.47	1	0.79	1	0.75	1
	260V	1.56	1	1.25	1	1.25	1
Slope of turn-off current dI_B/dt (normalised)	nom.	1	1	1	1	1	1
	190V	1	0.67	1	0.85	1	0.83
	260V	1	1.35	1	1.2	1	1.16
Excursion in base drive area of BU208A (note 2)	190V to 260V	 (note 3)		drive from SMPS *not* possible (note 4)			
Drive δ-range							
Allowed for deflection (Figs.8 and 9)	190V to 260V	0.36 – 0.7 (note 5)	0.35 – 0.72	0.36 – 0.72	0.35 – 0.72	0.36 – 0.7	0.35 – 0.72
Available from SMPS (Table 1)		0.43 – 0.63	0.37 – 0.57	0.3 – 0.39	0.61 – 0.7	0.39 – 0.48	0.52 – 0.61
Remarks		Note 3	Note 6	Note 4	Note 6		Note 6

1. Supply and deflection transistor operate in the simultaneous or the non-simultaneous mode.
2. Ref.6.
3. The large $I_{B\ end}$ variation caused by mains fluctuations must be stabilised.
4. Drive from SMPS not possible because available δ-range extends beyond allowable δ-range
5. With $I_{B\ end}$ stabilisation.
6. Precautions are required to prevent deflection transistor being forward biased during flyback period.

SUMMARY OF RECOMMENDATIONS

Where an SMPS is used in a tv receiver it should also preferably be used to supply the drive for the horizontal output transistor. This arrangement has several advantages:

- simpler filtering of the SMPS output as a result of the synchronisation of the SMPS and deflection circuit;
- a separate horizontal driver stage is not required;
- with a forward converter, a crowbar protection circuit is not required across the SMPS.

Complementary drive of the horizontal output stage is to be preferred since then the final value of the positive base current ($I_{B\ end}$) of the horizontal output transistor is not affected by changes in the duty factor of the SMPS.

Where mains isolation is not required, a forward-converter SMPS is preferred for two reasons:

- it imposes lower demands on the output circuit components (transistor, flywheel diode, choke, and output capacitor);
- it generates less interference than the flyback converter because the amplitude of the switching transients is smaller.

If mains isolation is essential, a flyback converter should be used.

REFERENCES

1. JANSSON, L.E., 'A survey of converter circuits for switched-mode power supplies', **Mullard Technical Communications**, Vol.12, No.119, July 1973, pp.271 to 278. Available as reprint Technical Note 24, TP1442/1.
2. DANGSCHAT, R., *et al*, 'Ein Sperrwandler-Netzmodul mit Netztrennung', Funkschau 47: 40-43, No.5, 1975.
3. WOLF, G., 'Mains isolating switch-mode power supply', **Electronic Applications Bulletin** 321, No.1, 1973, pp.28 to 43.
4. WHITE, L.M., 'Television switched-mode power supply using the TDA2640', **Mullard Technical Communications**, Vol.13, No.127, July 1975, pp.258 to 279.
5. JANSSON, L.E., 'Radio frequency interference suppression in switched-mode power supplies', **Mullard Technical Communications**, Vol.12, No.120, October 1973, pp.294 to 298. Available as reprint Technical Note 30, TP1443/1.
6. TEULING, D.J.A., *et al*, 'A self-regulating horizontal deflection circuit using transistors', **Electronic Applications Bulletin** 34, No.2, 1976, pp.78 to 89.
7. HETTERSCHEID, W. Th., 'Turn-on and turn-off behaviour of high-voltage switching transistors', **Electronic Applications Bulletin** 33, No.2, 1975, pp.59 to 72. See also: ARENTSEN, D. and SCHAIK, G. v., 'The effect of base-current waveform on the turn-off behaviour of high-voltage switching transistors', **Electronic Applications Bulletin** 34, No.2, pp.52 to 60.
8. 'High-voltage switching transistors BU207A, BU208A, and BU209A in the horizontal deflection circuit of CTV receivers', Technical Note 38, TP1546, Mullard Limited, 1975.

Product Information: Channel electron multipliers

INTRODUCTION

Electron multiplication by secondary emission within a tube of conducting glass has been a subject of intense study since the late '50s, and two related devices based on this phenomenon are now finding increasing application. The simplest, the single-channel electron multiplier (Fig. 1), is used mainly as a detector for charged particles, X-rays, and ultraviolet radiation. Its high electron gain, low background count rate, low power consumption, small size, and simplicity of use offer great advantages over the more complex conventional discrete-dynode multiplier. The second device of this type, the microchannel plate, is essentially an array of minute single-channel multipliers fused in the form of a disc. Its main application is in image intensification but it also can be used for radiation detection.

The single-channel multiplier and the microchannel plate are described briefly below; they are discussed in more detail in Refs. 1 and 2, copies of which are available on request.

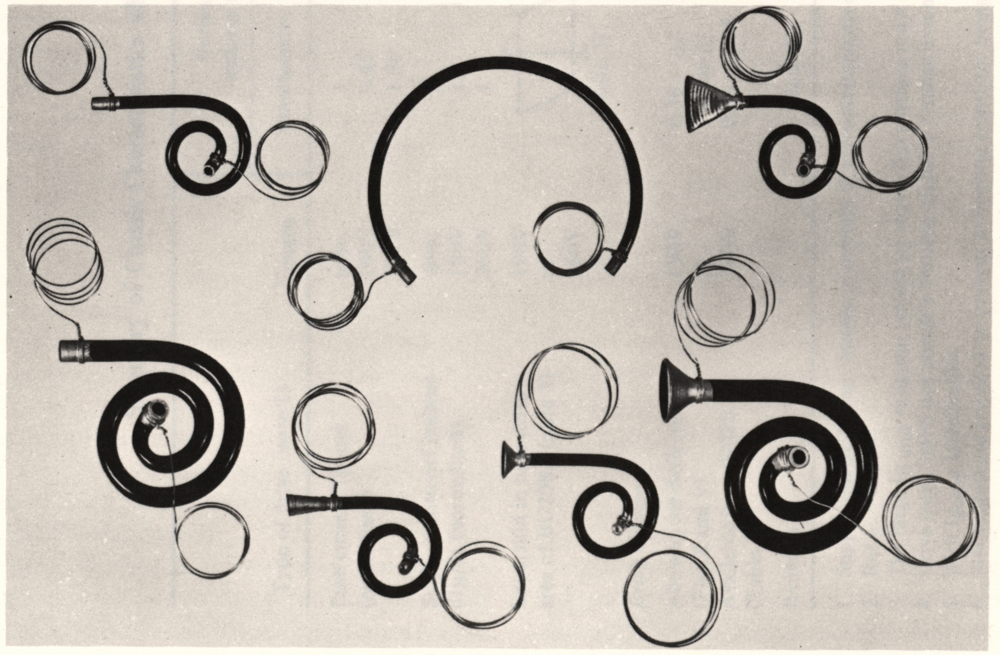


Fig. 1 — Single-channel electron multipliers

SINGLE-CHANNEL MULTIPLIERS

A single-channel electron multiplier consists of a hollow curved glass tube, with an internal diameter of a few millimetres and an inside resistive surface processed to have high secondary-emission coefficient. When a potential difference is applied between the ends of the tube, the resistive surface acts as a continuous dynode, electrically analogous to the separate dynodes of a conventional photomultiplier together with the resistor chain used to establish the separate dynode potentials. The device operates in vacuum.

When a charged particle or quantum of radiation enters the negative potential end of the tube, secondary electrons are generated and are accelerated along the tube until they strike the wall again, where they generate further secondary electrons. This avalanche process produces a large number of electrons at the positive end of the tube. This is illustrated in Fig. 2. The electrons appear as an output pulse that may contain more than 10^8 electrons. The pulse amplitude depends on the values of resistance and capacitance in the anode circuit of the multiplier.

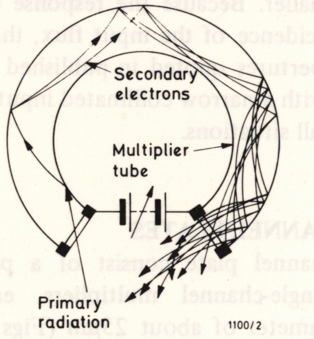


Fig.2 — Electron multiplication

Gain

The gain is an exponential and very steep function of voltage for values below 10^7 (see Fig.3). Above 10^7, saturation effects are observed which will be discussed later.

The multiplier will respond to ions, particles, X-rays, and other radiation. The detection efficiency varies for different forms of excitation, but any particle or quantum capable of exciting an electron from the dynode surface has a finite probability of detection.

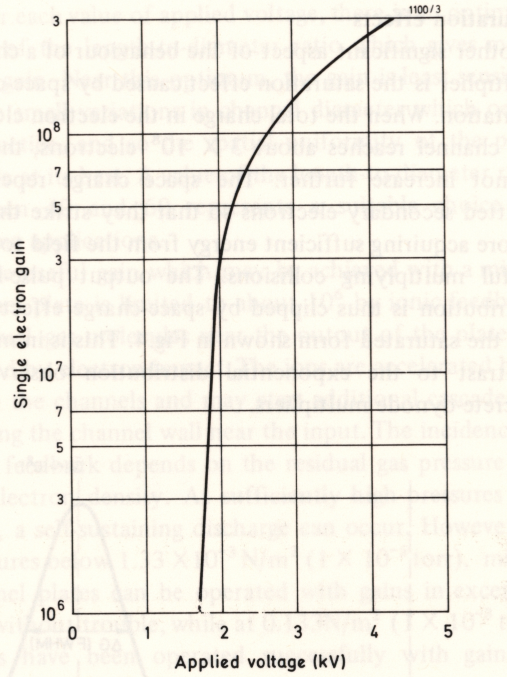


Fig.3 — Typical variation of gain with applied voltage

Since the resistive layer is continuous, many electron paths are possible, and the number of stages of multiplication is thus indeterminate. The electron trajectories are scaled in proportion to the dimensions of the channel for a given applied voltage. Thus if the length-to-diameter ratio is preserved, the same multiplication processes go on, and the same gain is achieved, irrespective of the absolute length of the channel. In practice it is necessary for the length-to-diameter ratio to exceed about 30:1. Channels are usually curved, and the gain is then less critically dependent on the length-to-diameter ratio. A ratio of about 60:1 is typical.

Ionic feedback

The tube forming the channel multiplier is curved because the gain of a straight tube would be sensitive to changes in vacuum pressure. In a straight tube, when the first cloud of electrons nears the output end of the multiplier, it is sufficiently dense to ionise a considerable number of the residual gas atoms in the tube. The positive ions drift under the influence of the applied field towards the more negative potential at the input end of the channel. The ions may acquire considerable energy before they collide with the wall of the tube. Consequently they may release electrons from the wall which will initiate a further avalanche, resulting in a spurious pulse. This process is repeated and thus a sequence of after-pulses may be observed, lasting approximately one microsecond, until the channel capacity is exhausted.

In a curved tube, the ions strike the wall of the tube before they have acquired sufficient energy to release secondary electrons. Electron multiplication is unaffected, since electrons need only to acquire an energy of about 50eV to release secondaries from the wall. The output of the curved multiplier is therefore virtually independent of ambient pressure, provided it does not exceed $6 \times 10^{-2}\,\text{Nm}^{-2}$. Above this pressure spurious pulses occur, and effects similar to those seen with straight channels are observed.

Saturation effects

Another significant aspect of the behaviour of a channel multiplier is the saturation effect caused by space-charge limitation. When the total charge in the electron cloud in the channel reaches about 3×10^8 electrons, the gain cannot increase further. The space charge repels the emitted secondary electrons so that they strike the wall before acquiring sufficient energy from the field to make useful multiplying collisions. The output pulse-height distribution is thus clipped by space-charge effects, and has the saturated form shown in Fig.4. This is in marked contrast to the exponential distribution observed in discrete-dynode multipliers.

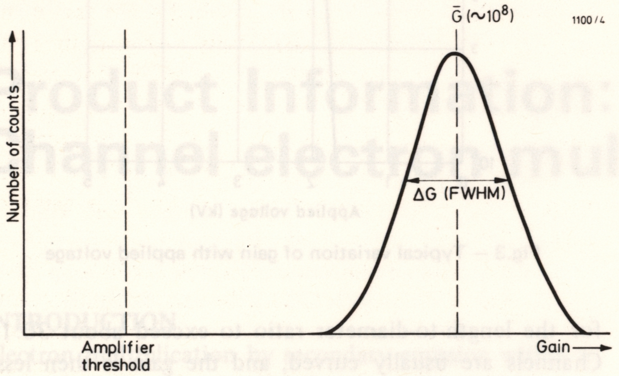


Fig.4 — Output pulse-height distribution

The resolution R of the pulse-height distribution is defined as:

$$R = \Delta G/\bar{G},$$

where ΔG is the full width at half maximum, and $\bar{G}$ is the modal value of the gain. The resolution is typically 30 to 60%. Furthermore, because the single-channel multiplier is inherently a high-gain device, the amplifier threshold can be set well below the modal gain level. Consequently, considerable gain fatigue can be tolerated in the multiplier before the gain falls to the point at which pulses are lost below the amplifier threshold.

When the multiplier is operated in the saturated mode described above, it is analogous to a Geiger counter, producing a pulse of given amplitude irrespective of the manner of excitation. It is unable in this condition to give information about the energy of the input stimulus, or about the number of particles should these enter the input simultaneously.

When the multiplier is not operated in its saturated mode, that is when the gain is less than 10^7, there is some proportionality between the input and output. However, there is a spread of pulse amplitudes because of the many possible electron paths through the multiplier. In this condition the pulse amplitude distribution is found to be approximately exponential, with smaller pulses being more probable than larger ones by an amount dependent on their amplitude.

As the channel multiplier curvature eliminates the possibility of after pulses, the dead time of the device is very short. However, if the mean pulse repetition rate is high, the field inside the channel is distorted. A state of dynamic equilibrium is then achieved, where the mean gain is reduced so that the average rate of flow of charge in the output pulses remains less than the charge that can be extracted from the resistive layer.

Input geometry

Larger input apertures can be achieved by producing multipliers with flared ends. The effective aperture (defined as the boundary within which the count rate does not fall below 50% of its maximum central value) is not necessarily the same as the geometrical aperture, but may be smaller. Because the response depends on the angle of incidence of the input flux, the dimensions of effective apertures quoted in published data, which are measured with a narrow collimated input beam, may not be valid in all situations.

MICROCHANNEL PLATES

A microchannel plate consist of a parallel array of straight single-channel multipliers, each having an internal diameter of about 25μm (Figs. 5 and 6). The plate is usually up to 70mm in diameter and 1mm thick. It is normally operated at about 1kV and has a gain of at least 10^3, typically 3×10^3.

Gain

A typical plot of current gain as a function of applied voltage for a microchannel plate is shown in Fig. 7. The channels have a length-to-diameter ratio of 40. With 0.5keV primary electrons at the input, the gain rises from about 10^3 at 900V to 1.5×10^5 at 1600V.

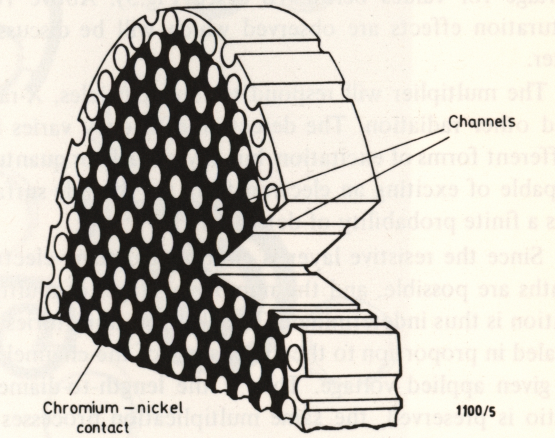


Fig. 5 — Schematic representation of microchannel plate (not to scale)

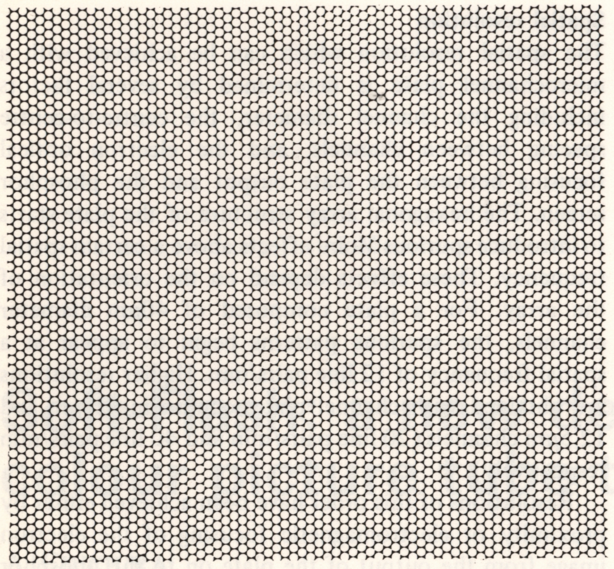


Fig. 6 — Magnified view of face of microchannel plate

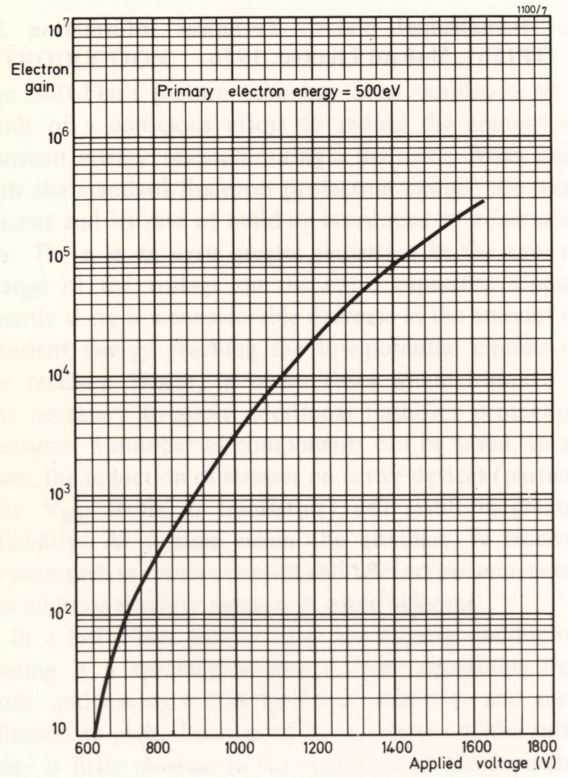


Fig. 7 — Current gain plotted against applied voltage

For each value of applied voltage, there is an optimum value of the length-to-diameter ratio which gives maximum gain. Near this optimum, the gain is least sensitive to the small variations in channel diameter which occur in practice, and so the spatial uniformity of the plate will be at its best. A value of the length-to-diameter ratio between 40 and 60 represents a suitable choice for imaging applications.

The useful gain which may be achieved with a microchannel plate is limited to about 10^6 by ionic feedback. Residual gas molecules near the output of the plate are ionised by electron impact. The ions are accelerated back down the channels and may start additional cascades by striking the channel wall near the input. The incidence of ionic feedback depends on the residual gas pressure and the electron density. At sufficiently high pressures and gains, a self-sustaining discharge can occur. However, at pressures below 1.33×10^{-3} N/m² (1×10^{-5} torr), microchannel plates can be operated with gains in excess of 10^5 without trouble, while at 0.133N/m² (1×10^{-3} torr), plates have been operated successfully with gains of several thousand.

In single-channel multipliers, ionic feedback effects are prevented by curving or spiralling the multiplier. Any ions formed can move only a short way down the multiplier before striking the wall. The energy on impact is scarcely sufficient to produce secondary electrons, and those that are generated can hardly multiply in the available length of channel. In these single multipliers, gains higher than 10^8 can be achieved before the gain is limited by space-charge effects. To obtain comparable gains with microchannel plates, two options are open. Either two plates can be operated in cascade, or a single plate with curved channels can be used.

Current transfer characteristics

The current transfer characteristics of a microchannel plate are shown in Fig. 8. With a resistance of 10^9 Ω and

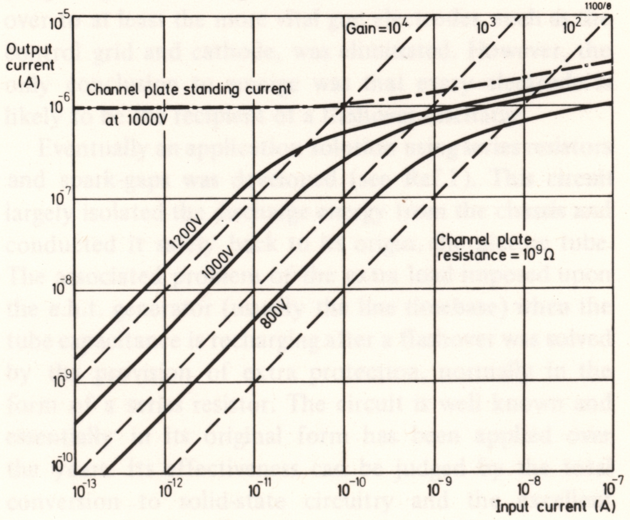


Fig. 8 — Current transfer characteristic

an operating potential of 1000V, the conduction current in the channel plate is 1μA. The transfer characteristic of the plate is linear for output currents up to 10% of the standing current. For higher input currents, the plate begins to show saturation effects, and the characteristic begins to level out. If the resistance of the plate remained constant, the output would saturate at a value close to the conduction current. However, the effective resistance of the plate changes slightly in operation because the cascade of electrons forms a parallel resistive path. This produces a more gradual saturation effect than might otherwise be expected, and allows an output current to be drawn which can exceed the nominal conduction current of 1μA. As the individual multipliers in the plate are virtually independent of one another, each channel will saturate if the current drawn from it approaches its standing current (that is, the current flowing in the channel wall).

The onset of saturation determines the maximum linear light output which can be obtained from a microchannel plate in combination with a phosphor screen at a particular accelerating voltage. For instance, with a typical phosphor efficiency of 20 lumens/watt and 8kV between the plate output face and the screen, the limit to the linear light output with a plate resistance of $10^9\Omega$ is approximately 10.3cd/m^2. Acceptable images with some loss of contrast can be obtained at higher light levels. As microchannel plates can be made with various resistances, the saturation characteristics can be chosen to suit the application, provided the power dissipated in the plate is not excessive.

Spatial resolution

The spatial resolution which can be achieved with a microchannel plate is determined by several factors including the channel diameter, the centre-to-centre spacing between channels, and the electron optical focusing between the plate output and the phosphor screen. The hexagonal array structure of the plate starts to introduce spurious Moiré effects above a certain spatial frequency. These effects depend on the orientation of the array. For a plate composed of 40μm diameter channels, the spurious patterns begin to appear at a spatial frequency of 12 line pairs/mm. Higher resolutions are obtained with channel plates comprising 25μm and 12.5μm channels (20 line pairs/mm and 40 line pairs/mm respectively).

Resolutions of this order have been achieved either by proximity or electromagnetic focusing of the electron image from the output of the plate on to the phosphor screen.

REFERENCES

1. 'Single-channel electron multipliers', Technical Information 16, TP1504, Mullard Limited, 1975.
2. 'Microchannel plates', Technical Information 31, TP1561, Mullard Limited, 1976.

Soft-Flash: a new development in colour picture tube technology

A. CIUCIURA
Mullard Application Laboratory

The Soft-Flash tv picture tube giving greatly improved flashover protection is described. A built-in internal resistance reduces both the peak flashover current and the rate of current increase by a factor of 10, and the rate of current fall by a factor of 100.

INTRODUCTION

The Soft-Flash picture tube has been developed as a result of a conscious effort to reduce the amount of transient energy released during a flashover. When used with the standard flashover protection circuit, the peak current and its rate of build-up is reduced by a factor of ten. There is an even greater reduction in the rate of change of the voltage on the e.h.t. capacitor. Consequently there is a considerable decrease in the amount of transient energy reaching the low-potential circuits of the receiver. Hence, in many cases where hitherto it was necessary to apply additional flashover protection measures, a number of components can be saved. In all cases, the reduction of stresses on active devices (particularly V_{BE} levels in transistors) will result in greater reliability. At a time when the auxiliary tv receiver services such as remote control and teletext are increasing, this additional safety measure is most welcome.

In a Soft-Flash picture tube, the internal conductive coating is a specially-developed layer containing iron oxide and having a high electrical resistivity and good adhesion to glass. Because of the geometry of the tube, there is little increase in the resistance of the capacitor formed by the inner and outer conductive coatings, but the effective resistance to currents flowing through the gun structure during a flashover is increased to about 400Ω. This increase in resistance produces a marked decrease in the amount of transient energy reaching the low-voltage circuits of the receiver during a flashover.

FLASHOVER

The problems of picture tube flashover and its possible adverse affects on receiver reliability are well known. Some ten years ago, when serious thought was given to the design of the first solid-state receivers, these problems seemed almost insuperable. Then Mullard launched a programme of work aiming at a solution. At an early stage, it was shown that no matter how well a picture tube is designed or how carefully it is made, sooner or later it will produce a flashover. Attempts were made to design a picture tube in which the possibility of flashover to at least the more vital gun electrodes, such as the control grid and cathode, was eliminated. However, the only conclusion to emerge was that every electrode is likely to be the recipient of a flashover discharge.

Eventually an application solution using series resistors and spark-gaps was developed (see Ref.1). This circuit largely isolated the discharge energy from the chassis and conducted it safely back to its origin, the picture tube. The associated problem of the extra load imposed upon the e.h.t. generator (usually the line timebase) when the tube capacitance is recharging after a flashover was solved by the provision of extra protection, normally in the form of a series resistor. The circuit is well known and essentially in its original form has been applied over the years. Its effectiveness can be judged by the total conversion to solid-state circuitry and the excellent reliability of modern television receivers.

While this approach was a great advance, a small fraction of the transient energy generated by a flashover still penetrates the low-potential circuits. In the past, this has resulted in failures of scan rectification diodes, and caused problems with tuners and thyristor-controlled power supplies. It has also sometimes resulted in the unwarranted activation of overvoltage protection circuits. With the modern trend towards the introduction of a multiplicity of complex integrated circuits into tv receivers (for example LSI, remote control, and teletext), there have been some forebodings that these difficulties might increase.

The key to the solution was found during the original flashover research programme. It was observed that transients resulting from a flashover could be considerably reduced by the provision of an internal resistor inside the picture tube. A great deal of work was necessary before this could be achieved but, after some research, a practical design was developed. As a consequence, picture tubes, interchangeable with existing tubes, having a new internal conductive layer with considerably increased resistance, are now being mass-produced.

EVALUATION OF A SIMPLE MODEL

The component parts of a flashover circuit can be considered as a series circuit containing distributed components L, C, and R (see Fig.1). The voltages and currents in this circuit during a flashover can be calculated using a very simple model consisting of lumped components (see Appendix and Ref.1). The resulting waveforms for a standard 26in tube (internal resistance about 20Ω) and a Soft-Flash tube (internal resistance about 400Ω) are compared in Figs.2 and 3. To assist comparison, the vertical and horizontal scales in both figures are the same.

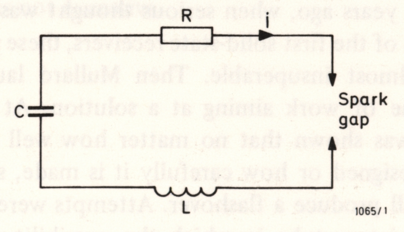


Fig.1 — Simplified flashover circuit

The differences are striking: while a standard tube is represented by a circuit with damped oscillations, the Soft-Flash tube is shown as an over-damped circuit. The latter has a peak current which is about one tenth of that in the standard tube. However, the area under the current curve for both figures is the same (assuming that the picture tube capacitance is completely discharged by a flashover). The difference in the decay of current is even more pronounced; in terms of fall-time, it is 0.1μs for the standard tube and 2.1μs for the Soft-Flash tube. Associated with it is the decay of voltage on the picture tube capacitance. While the standard tube has a negative overswing, the Soft-Flash tube has a comparatively slow decay with a time-constant of approximately CR. This implies that the transfer of energy from the internal conductive coating to the deflection coils (or any other part of the circuit) is at a much slower rate.

PRACTICAL SOFT-FLASH TUBE

The above considerations indicate that it would be advantageous to increase the series resistance of the flashover circuit. Theoretically this added resistance could be placed anywhere. External to the tube, however, the only accessible location is in the connecting lead between the tube base and the external Aquadag coating. As shown in Fig.3, the voltage developed across this resistance during flashover reaches e.h.t. level, a disturbance which would create even more problems than the resistor would solve. Thus, the only feasible position for the resistor is inside the tube, preferably between the e.h.t. capacitor and the final anode.

The specifications for such a resistor are stringent: firstly, it must be capable of handling high voltages (e.h.t. plus normal tolerances); secondly, its reliability must be such as not to reduce the life of the tube; and thirdly, it must not be affected by tube processing, and in turn it must not affect the normal operation of the tube in any way.

Of several possible approaches, it was decided that an increase in the resistivity of the internal conductive layer was the most promising. A long period of research followed, leading to the development of a new material of suitable composition. This material allows tight control of resistivity, has good adhesion to glass, can be deposited in a controlled way, and does not contain any elements which could affect the internal chemistry of the tube.

The new inside coating is readily recognisable by reddish colouring. However, it should not be automatically assumed that every tube with a reddish internal coating is a Soft-Flash tube.

PERFORMANCE

The waveform of current produced by an internal flashover in a standard tube is shown in Fig.4, and by a flashover in a Soft-Flash tube in Fig.5. Bearing in mind the great simplicity of the model described earlier, the similarity between these oscillograms and Figs.2 and 3 is quite good. The expected reduction in the peak current has been realised and the general shapes of the waveforms are of the same order. The rate of rise of current for the Soft-Flash tube is about one-tenth that in the standard tube which implies that the transients produced across

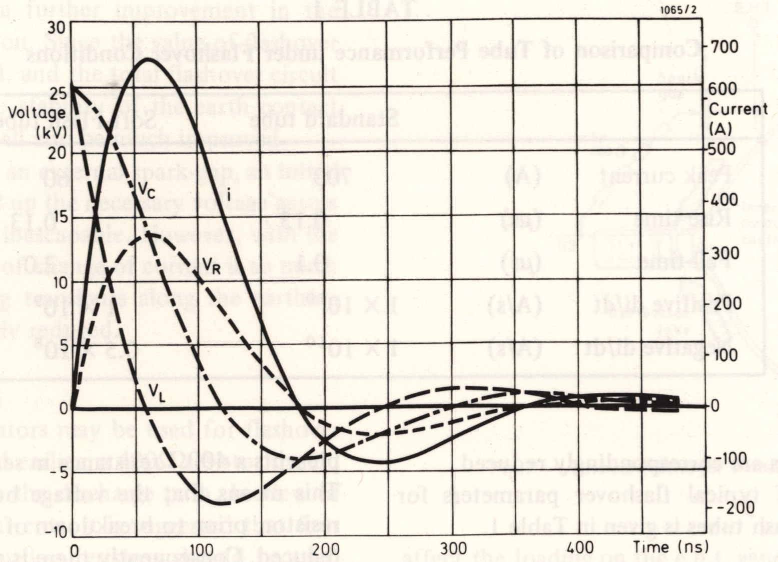


Fig.2 — Flashover waveforms in simplified model of standard tube

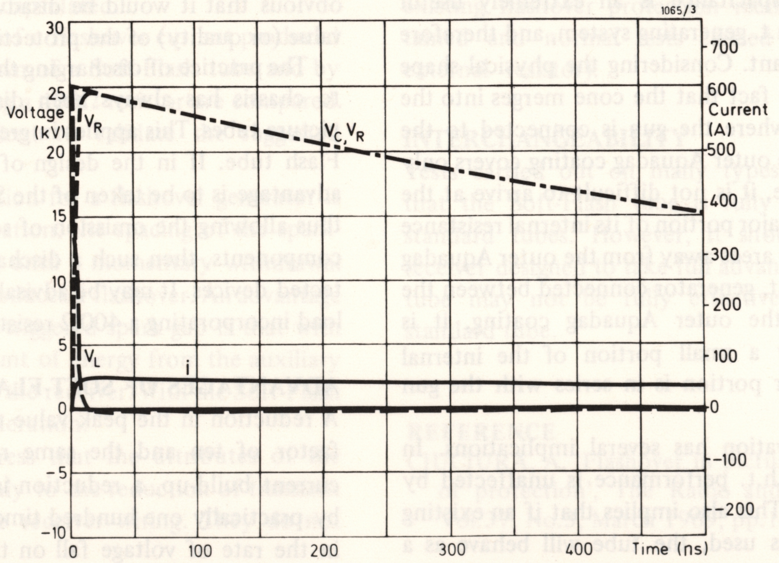


Fig.3 — Flashover waveforms in simplified model of Soft-Flash tube

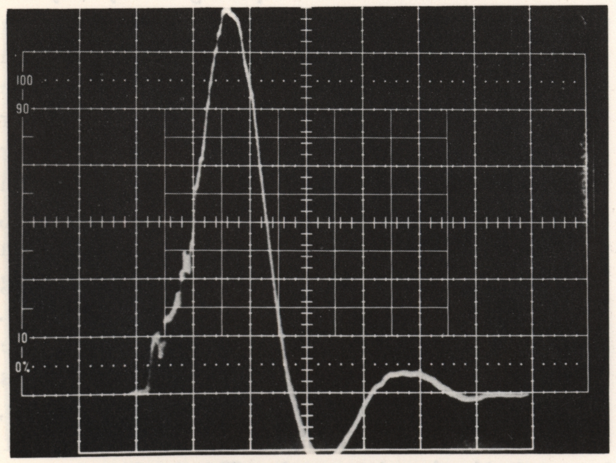


Fig.4 — Flashover current in standard tube

Vertical scale : 100A/division Horizontal scale : 100ns/division

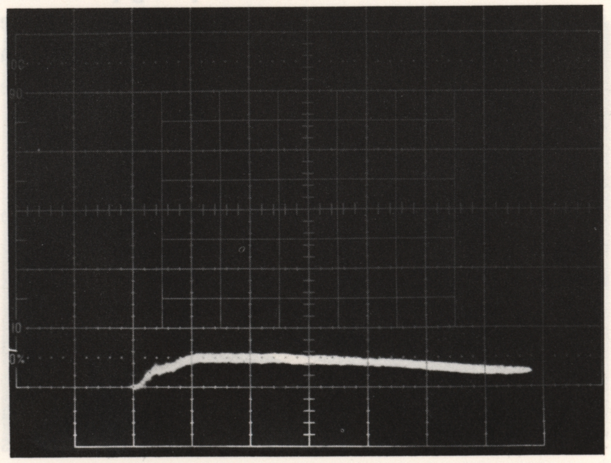


Fig.5 — Flashover current in Soft-Flash tube

Vertical scale : 100A/division Horizontal scale : 100ns/division

TABLE 1

Comparison of Tube Performance under Flashover Conditions

		Standard tube	Soft-Flash tube
Peak current	(A)	700	60
Rise time	(μs)	0.13	0.13
Fall-time	(μs)	0.1	2.0
Positive di/dt	(A/s)	1×10^{10}	1×10^{9}
Negative di/dt	(A/s)	1×10^{10}	0.5×10^{8}

inductive components are correspondingly reduced.

A comparison of typical flashover parameters for standard and Soft-Flash tubes is given in Table 1.

SOFT-FLASH TUBE AS E.H.T. CAPACITOR

The picture tube capacitance is an extremely useful component of the e.h.t. generating system, and therefore its quality is important. Considering the physical shape of this capacitor, the fact that the cone merges into the neck of the tube (where the gun is connected to the inside layer), and the outer Aquadag coating covers only a portion of the cone, it is not difficult to arrive at the conclusion that the major portion of its internal resistance is concentrated in the areas away from the outer Aquadag coating. With the e.h.t. generator connected between the e.h.t. contact and the outer Aquadag coating, it is presented with only a small portion of the internal resistance. The major portion is in series with the gun structure.

The above observation has several implications. In the first place the e.h.t. performance is unaffected by the Soft-Flash tube. This also implies that if an existing flashover generator is used, the tube will behave as a normal tube. For this reason, a 400Ω resistor should be connected in series with the main spark-gap. Another implication is that during a flashover inside the picture tube a 400Ω resistor is present in series with the gun structure, that is in series with the flashover discharge path.

FLASHOVER PROTECTION

A peak current of 60A is obviously too high to be discharged directly into receiver circuitry, especially bearing in mind that practically the whole charge stored in the picture tube capacitance is removed by a flashover. Thus the flashover protection circuit as recommended for a standard tube should be retained.

In fact it is not possible to suggest any reduction in the values of protective resistors, particularly the resistors in series with the cathode pins. As mentioned in the previous section, during a flashover the Soft-Flash tube presents a 400Ω resistance in series with the e.h.t. source. This means that the voltage build-up on the protective resistor, prior to breakdown of the external spark-gap, is reduced. Consequently there is a slight delay in the firing of the spark-gap. While this by itself is not harmful, it is obvious that it would be disadvantageous to reduce the value (or quality) of the protection resistors.

The practice of discharging the e.h.t. capacitor directly to chassis has always been discouraged with standard picture tubes. This applies in greater measure to the Soft-Flash tube. If in the design of the receiver circuits full advantage is to be taken of the Soft-Flash characteristics, thus allowing the omission of some flashover protection components, then such a discharge can endanger unprotected devices. It may be advisable to include a discharge load incorporating a 400Ω resistor.

ADVANTAGES OF SOFT-FLASH TUBE

A reduction in the peak value of flashover current by a factor of ten and the same reduction in the rate of current build-up, a reduction in the rate of current fall by practically one hundred times and a similar reduction in the rate of voltage fall on the e.h.t. capacitor, these are the main attributes of the Soft-Flash tube. All this has been obtained without impairing in any way the normal electrical, mechanical, or optical characteristics.

In practical terms this implies a considerable reduction in the amount of transient energy which bypasses the normal protective measures and is injected into receiver circuitry. One of the more obvious paths for the transfer of energy is from the internal coating into the deflection coil system and thus into the line output transformer, affecting a number of diodes (such as scan rectification diodes). There is always some capacitive coupling between the picture tube and the rest of the receiver. However, with the slowing down of the rate of the voltage change on the picture tube capacitance, this coupling will be reduced substantially. Furthermore, considering the flashover path, the major resistive element is now inside the picture tube. This implies that the voltage drop on the outside Aquadag coating is reduced by a factor of ten – a further improvement in the transient coupling situation. Since the value of flashover current has been reduced, and the total flashover circuit impedance increased, the stability of the earth contact to the Aquadag coating will also be much improved.

For the breakdown of an external spark-gap, an initial inrush of current to build-up the necessary voltage across the protective resistor is inescapable. However, with the Soft-Flash tube the rate of change of current is so much slower that the resulting transients along the earthing system are correspondingly reduced.

FLASHOVER TESTING

Existing flashover generators may be used for flashover testing, but as mentioned earlier, a 400Ω resistor should be inserted in series with the discharge path. In view of the high voltages and currents, it is suggested that four 100Ω 1W carbon-composition resistors should be used for this purpose, well separated from the rest of the wiring, and preferably encapsulated.

In Fig.6 oscillograms of a flashover current produced by the internal breakdown of a Soft-Flash tube and by a flashover generator using a 400Ω resistor are compared. The similarity is sufficient to validate the suggested approach.

An interesting suggestion for a flashover generator is shown in Fig.7. In operation, the spacing of the spark-gap is gradually reduced until a momentary withdrawal of the insulating wafer produces a flashover. An advantage of this approach over a triggered spark-gap is that with the latter a certain amount of energy from the auxiliary spark-gap is injected into the receiver. With the Soft-Flash tube this may be a consideration.

It is necessary to stress that the attributes of the Soft-Flash tube apply only to the reduction of transient energy injected into the receiver wiring. They do not affect the loading on the e.h.t. generator imposed by the recharge of the e.h.t. capacitor after a flashover. The existing flashover protection method should be maintained and normal tests carried out (using a 400Ω external resistor).

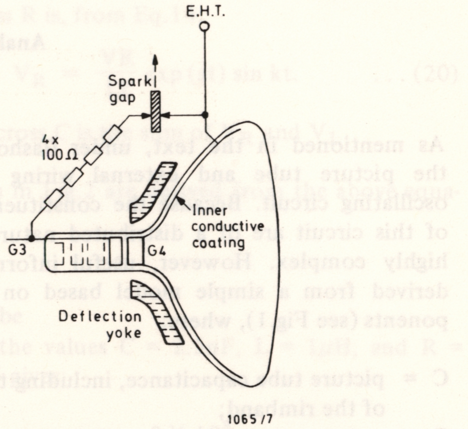


Fig.7 – A novel flashover simulation circuit

INTERCHANGEABILITY

Tests carried out on many types of receiver indicate that the Soft-Flash tube is fully interchangeable with standard tubes. However, it should be noted that a receiver designed to take full advantage of the Soft-Flash tube may not be fully effective when fitted with a standard tube.

REFERENCE

CIUCIURA, A. 'Flashover in picture tubes and methods of protection', The Radio and Electronic Engineer, Vol.37, No.3, March 1969, pp.149 to 168.

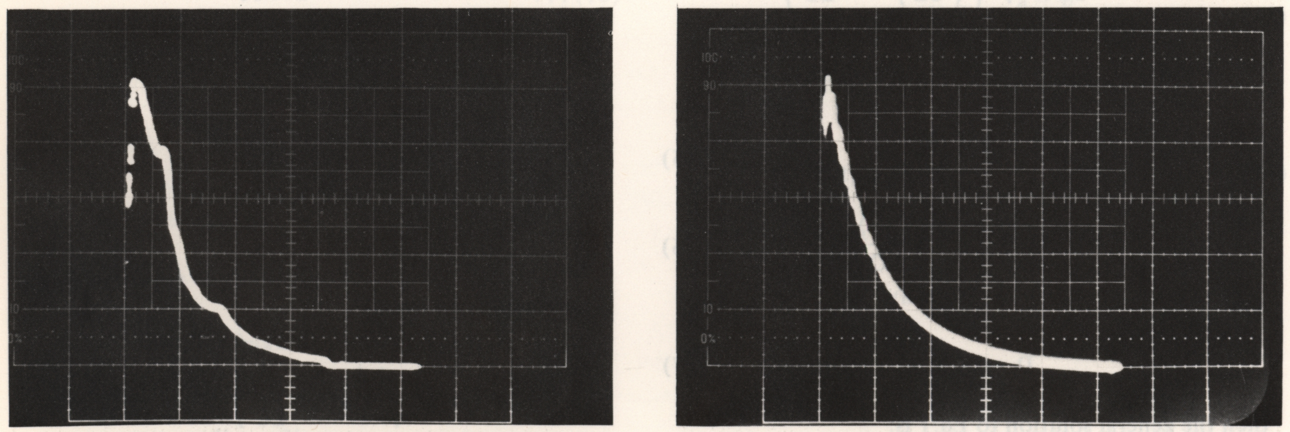


Fig.6 – (a) Flashover current in Soft-Flash tube. (b) Simulated flashover current
Vertical scale: 10A/division. Horizontal scale: 1μs/division

APPENDIX

Analysis of Simple Model of Flashover Circuit

As mentioned in the text, under flashover conditions the picture tube and external wiring form a series oscillating circuit. Because the constituent components of this circuit are of a distributed nature the circuit is highly complex. However, useful information can be derived from a simple model based on lumped components (see Fig.1), where:

C = picture tube capacitance, including the capacitance of the rimband;
R = series resistance, primarily of the internal and external picture tube coatings;
L = series inductance, primarily of the external connecting wires and the gun structure;
G = interelectrode spark gap.

For a standard 26in colour tube operating at an applied voltage of 25kV, C = 2.5nF, L = 1μH, and R = 20Ω. For a Soft-Flash tube the values are the same except for R = 400Ω.

Consider time t = 0 at which C is charged to 25kV, G is closed, and current i is zero. The discharge loop is represented by the equation:

$$\mathrm{Ri} + \mathrm{L}\frac{\mathrm{di}}{\mathrm{dt}} + \frac{1}{\mathrm{C}}\int \mathrm{i\,dt} = 0. \qquad \ldots(1)$$

Differentiating and rearranging gives:

$$\frac{\mathrm{d^2i}}{\mathrm{dt^2}} + \frac{\mathrm{R}}{\mathrm{L}}\frac{\mathrm{di}}{\mathrm{dt}} + \frac{1}{\mathrm{LC}} = 0. \qquad \ldots(2)$$

Using the auxiliary equation:

$$\mathrm{D}^2 + \frac{\mathrm{R}}{\mathrm{L}}\mathrm{D} + \frac{1}{\mathrm{LC}} = 0, \qquad \ldots(3)$$

whose roots are:

$$\mathrm{D} = -\frac{\mathrm{R}}{2\mathrm{L}} \pm \sqrt{\left\{\left(\frac{\mathrm{R}}{2\mathrm{L}}\right)^2 - \frac{1}{\mathrm{LC}}\right\}}\,; \qquad \ldots(4)$$

and taking:

$$\mathrm{a} = -\frac{\mathrm{R}}{2\mathrm{L}}, \qquad \ldots(5)$$

$$\mathrm{b} = \sqrt{\left(\mathrm{a}^2 - \frac{1}{\mathrm{LC}}\right)}, \qquad \ldots(6)$$

and:

$$\mathrm{D = a \pm b;} \qquad \ldots(7)$$

then the general solution to Eq.1 is:

$$\mathrm{i = A\exp(a+b)t + B\exp(a-b)t.} \quad \ldots(8)$$

Since at t = 0, i = 0, then, from Eq.8:

$$\mathrm{A + B = 0.} \qquad \ldots(9)$$

Differentiating Eq.8 gives:

$$\frac{\mathrm{di}}{\mathrm{dt}} = \mathrm{A(a+b)\exp(a+b)t + B(a-b)\exp(a-b)t.} \qquad \ldots(10)$$

At t = 0, i = 0 and all the voltage on C appears across L. Thus:

$$\mathrm{L}\frac{\mathrm{di}}{\mathrm{dt}} = \mathrm{V},$$

or:

$$\frac{\mathrm{di}}{\mathrm{dt}} = \frac{\mathrm{V}}{\mathrm{L}}.$$

Substituting in Eq.10 $\frac{\mathrm{di}}{\mathrm{dt}} = \frac{\mathrm{V}}{\mathrm{L}}$ and t = 0 gives:

$$\frac{\mathrm{V}}{\mathrm{L}} = \mathrm{Aa + Ab + Ba - Bb.} \qquad \ldots(11)$$

Combining Eqs.10 and 11 and solving gives:

$$\mathrm{A} = \frac{\mathrm{V}}{2\mathrm{Lb}} \text{ and } \mathrm{B} = \frac{-\mathrm{V}}{2\mathrm{Lb}}.$$

Substituting for A and B in Eq.8 gives the complete solution to Eq.1; that is:

$$\mathrm{i} = \frac{\mathrm{V}}{2\mathrm{Lb}}\Big[\mathrm{\exp(a+b)t - \exp(a-b)t}\Big] \quad \ldots.(12)$$

Standard tube

Substituting the values given earlier for a standard tube (that is, C = 2.5nF, L = 1μH, and R = 20Ω) in Eq.5:

$$\mathrm{a} = -\frac{\mathrm{R}}{2\mathrm{L}} = -\frac{20}{2 \times 10^{-6}} = -10^7,$$

and:

$$\mathrm{a}^2 = 10^{14}.$$

Also:

$$\frac{1}{\mathrm{LC}} = \frac{1}{10^{-6} \times 2.5 \times 10^{-9}},$$

$$= 4 \times 10^{14}.$$

Therefore, from Eq.6:

$$\mathrm{b} = \sqrt{\left(\mathrm{a}^2 - \frac{1}{\mathrm{LC}}\right)},$$

$$= 10^7\sqrt{(-3)} = 10^7\,\mathrm{j}\sqrt{3}.$$

Let b = jk; then:

$$\mathrm{k} = 10^7\sqrt{3}.$$

Substituting b = jk in Eq.12 and rearranging gives:

$$\mathrm{i} = \frac{\mathrm{V}}{\mathrm{Lk}}\exp(\mathrm{at})\left[\frac{\exp(\mathrm{jkt}) - \exp(-\mathrm{jkt})}{2\mathrm{j}}\right],$$

or: ...(13)

$$\mathrm{i} = \frac{\mathrm{V}}{\mathrm{Lk}}\exp(\mathrm{at})\sin(\mathrm{kt}). \qquad \ldots(14)$$

Substituting values for V, L, k, and a in Eq.14 gives:

$$\mathrm{i} = \frac{2.5 \times 10^3}{\sqrt{3}}\exp(-10^7\mathrm{t}) \cdot \sin(10^7\sqrt{3}\mathrm{t}). \qquad \ldots(15)$$

The rate of change of current is obtained by differentiating Eq.14; that is:

$$\frac{\mathrm{di}}{\mathrm{dt}} = \frac{\mathrm{V}}{\mathrm{Lk}}\exp(\mathrm{at})\Big\{\mathrm{a}\sin(\mathrm{kt}) + \mathrm{k}\cos(\mathrm{kt})\Big\}. \qquad \ldots(16)$$

Let $\tan\alpha = \frac{\mathrm{a}}{\mathrm{k}}$ so that $\mathrm{a} = \sqrt{(\mathrm{a}^2 + \mathrm{k}^2)} \cdot \sin\alpha$ and

$$\mathrm{k} = \sqrt{(\mathrm{a}^2 + \mathrm{k}^2)} \cdot \cos\alpha.$$

Then:

$$\frac{\mathrm{di}}{\mathrm{dt}} = \frac{\mathrm{V}}{\mathrm{Lk}}\exp(\mathrm{at})\sqrt{(\mathrm{a}^2 + \mathrm{k}^2)} \cdot \cos(\mathrm{kt} - \alpha). \qquad \ldots(17)$$

The maximum value of i occurs when:

$$\mathrm{kt} - \alpha = \frac{\pi}{2};$$

that is, at the time given by:

$$\mathrm{t_{i(max)}} = \frac{\pi}{2\mathrm{k}} + \frac{\alpha}{\mathrm{k}} = \frac{1}{\mathrm{k}}\left\{\frac{\pi}{2} + \tan^{-1}\left(\frac{\mathrm{a}}{\mathrm{k}}\right)\right\}. \qquad \ldots(18)$$

Substituting for k and a gives:

$$\mathrm{t_{i(max)}} = \frac{1}{10^7 \times \sqrt{3}}\left\{\frac{\pi}{2} + \tan^{-1}\left(\frac{-10^7}{10^7\sqrt{3}}\right)\right\},$$

$$= 60.46\mathrm{ns};$$

and substituting this value in Eq.15 gives:

$$\mathrm{i_{max}} = 682\mathrm{A}.$$

The voltage developed across L is, from Eq.17:

$$\mathrm{V_L} = \mathrm{L}\frac{\mathrm{di}}{\mathrm{dt}} = \frac{\mathrm{V}}{\mathrm{k}}\exp(\mathrm{at})\cos(\mathrm{kt} - \alpha); \qquad \ldots(19)$$

and that across R is, from Eq.14:

$$\mathrm{V_R} = \frac{\mathrm{VR}}{\mathrm{Lk}}\exp(\mathrm{at})\sin\mathrm{kt}. \qquad \ldots(20)$$

The voltage across C is the sum of $\mathrm{V_R}$ and $\mathrm{V_L}$.

The curves in Fig.2 are derived from the above equations.

Soft-Flash tube

Substituting the values C = 2.5nF, L = 1μH, and R = 400Ω in Eq.5 gives:

$$\mathrm{a} = -2 \times 10^8,$$

and:

$$\mathrm{a}^2 = 4 \times 10^{16},$$

and, from Eq.6:

$$\mathrm{b} = \sqrt{(3.96)} \times 10^8.$$

Since b is real, Eq.12 applies directly; thus, substituting the appropriate values gives:

$$\mathrm{i} = 62.814\,\Big\{\exp(-1.0025 \times 10^6\mathrm{t}) - \exp(-3.99 \times 10^8\mathrm{t})\Big\}. \qquad \ldots(21)$$

Differentiating Eq.12 gives:

$$\frac{\mathrm{di}}{\mathrm{dt}} = \frac{\mathrm{V}}{2\mathrm{Lb}}\Big\{\mathrm{(a+b)\exp(a+b)t - (a-b)\exp(a-b)t}\Big\}. \qquad \ldots(22)$$

The maximum value of i occurs when:

$$\mathrm{(a+b)\exp(a+b)t = (a-b)\exp(a-b)t,}$$

or:

$$\exp(2\mathrm{bt}) = \frac{\mathrm{a-b}}{\mathrm{a+b}}. \qquad \ldots(23)$$

The maximum current occurs at $\mathrm{t_{i(max)}}$ given by:

$$\mathrm{t_{i(max)}} = \frac{\log_e\left(\frac{\mathrm{a-b}}{\mathrm{a+b}}\right)}{2\mathrm{b}}. \qquad \ldots(24)$$

Substituting for a and b in Eq.24 gives:

$$\mathrm{t_{i(max)}} = 15.04\mathrm{ns},$$

and:

$$\mathrm{i_{max}} = 61.72\mathrm{A}.$$

These values are considerably less than those for a standard tube (60.46ns and 682A respectively).

The voltage developed across L is, from Eq.22:

$$V_L = L\frac{di}{dt},$$

$$= \frac{V}{2b}\left[(a+b)\exp(a+b)t - (a-b)\exp(a-b)t\right], \qquad \ldots(25)$$

and the voltage developed across R is, from Eq.12:

$$V_R = \frac{VR}{2Lb}\left[\exp(a+b)t - \exp(a-b)t\right]. \qquad \ldots(26)$$

The voltage developed across C is the sum of V_R and V_L.

The curves in Fig.3 are derived from the above equations.

It is of interest to note that for a Soft-Flash tube:

$$a^2 = 4 \times 10^{16} \text{ and } 1/LC = 4 \times 10^{14};$$

that is:

$$a^2 \gg \frac{1}{LC}. \qquad \ldots(27)$$

Now, from Eq.6:

$$b = \sqrt{\left(a^2 - \frac{1}{LC}\right)} = a\sqrt{\left(1 - \frac{1}{a^2LC}\right)}.$$

Using the binomial expansion:

$$b \simeq a\left(1 - \frac{1}{2a^2LC}\right).$$

Thus:

$$a + b = a + a - \frac{a}{2a^2LC} = 2a - \frac{1}{2aLC}.$$

But, from Eq.27:

$$|a| \gg \left|\frac{1}{aLC}\right|.$$

Therefore:

$$a + b \simeq 2a = -\frac{R}{L},$$

and:

$$a - b = a - a + \frac{a}{2a^2LC} = \frac{1}{2aLC} = -\frac{1}{RC}.$$

Also, the term 2Lb in Eq.12 becomes:

$$2Lb = 2La\left(1 - \frac{1}{2a^2LC}\right),$$

$$= 2La - \frac{1}{aC}.$$

But, from Eq.27:

$$|La| \gg \left|\frac{1}{aC}\right|.$$

Therefore:

$$2Lb = 2La = -R.$$

Substituting these values in Eq.12 gives:

$$i = -\frac{V}{R}\left\{\exp\left(-\frac{R}{L}t\right) - \exp\left(-\frac{1}{CR}t\right)\right\}. \qquad \ldots(28)$$

Substituting for a^2 in Eq.27 gives:

$$\frac{R^2}{4L^2} \gg \frac{1}{LC} \text{ or } \frac{R}{4L} \gg \frac{1}{CR}$$

Thus for small values of t, when the current is increasing, the current build-up is controlled by the time-constant L/R and is approximated by:

$$i_b = \frac{V}{R}\left\{1 - \exp\left(-\frac{R}{L}t\right)\right\}. \qquad \ldots(29)$$

Similarly, the decay of current is controlled by the time-constant CR and is approximated by:

$$i_d \doteqdot \frac{V}{R}\exp\left(-\frac{t}{CR}\right). \qquad \ldots(30)$$

Under these conditions, the maximum value of current is simply:

$$i_{max} = \frac{V}{R}.$$

The maximum time at which the maximum current occurs is found by differentiating Eq.28 and equating to zero; that is:

$$\frac{1}{CR}\exp\left(-\frac{t}{CR}\right) = \frac{R}{L}\exp\left(-\frac{R}{L}t\right),$$

or, since $\frac{R}{L} \gg \frac{1}{CR}$:

$$\frac{CR^2}{L} = \exp\left(-\frac{t}{CR} + \frac{R}{L}t\right) \simeq \exp\left(\frac{R}{L}t\right),$$

and:

$$t_{max} = \frac{L}{R}\log_e\left(\frac{C}{L}R^2\right).$$